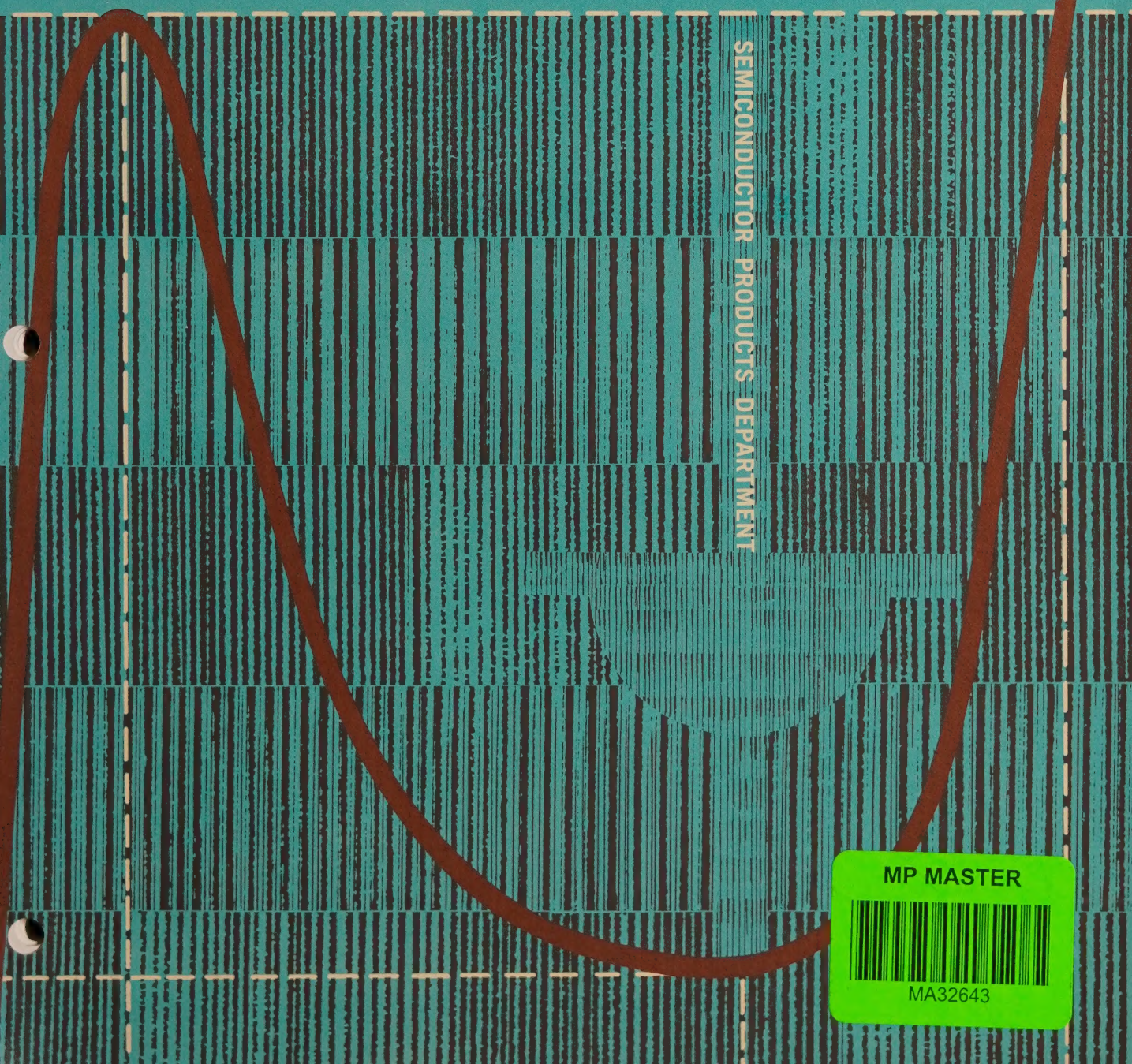
TUNNEL DIODE

APPLICATION NOTES

SEMICONDUCTOR PRODUCTS DEPARTMENT



MP MASTER
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A SURVEY OF UHF AND MICROWAVE TUNNEL DIODE APPLICATIONS

by

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In September of 1959, General Electric introduced TO-18 packaged tunnel diodes with resistive cut-off frequencies in the 2 - 3 KMC/s ranges. These units led to many of the investigations reported in this survey. As a result of this activity, improved microwave packaged tunnel diodes have made their appearance on the market in recent months. General Electric introduced the 1N3218, 1N3218A, 1N3219, and 1N3219A microwave units in a small stripline package having a total typical inductance of only 0.3 nanohenries. These units have cut-off frequencies in the 3 - 5 KMC/s region. (For individual data, see specification sheets on pages 86 and 87 of this report).

It is the purpose of this survey to present:

a. A concentrated Bibliography of some pertinent papers.

b. A short summary of each paper highlighting the key points and the author's conclusions.

c. Comments on the significance of the results achieved or the methods used by the author.

d. Recommend a G. E. tunnel diode suitable for the application.

Introduction:

This survey deals with papers published over the last year on the subject of tunnel diode useage at UHF and microwave frequencies. It is subdivided into four basic sections:

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Section 2)	Oscillators	36-55
Section 3)	Converters: a. driven mixers b. autodyne converters	56-84

BIBLIOGRAPHY

Section I - Amplifiers

1) Yariv A, Cook, J.S. "A Noise Investigation of Tunnel Diode Microwave Amplifiers".
IRE Proceedings, April 1961, pp. 739-743.

2) Yariv, A., Cook, J.S. and Butzien, P.E. "Operation of an Esaki Diode Microwave Amplifier".
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G. E. Technical Information Series #R60 ELS-102, Oct. 1960.

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SECTION I

AMPLIFIERS

ABSTRACT:

The author's of the following articles have taken different approaches to the tunnel diode - UHF or microwave - amplifier design. Each has considered somewhat different design aspects and/or methods. If one compiles their individual recommendations, however, one can draw a number of important conclusions - both from a device and circuit standpoint.

a. Gain:

To obtain high gain one must make the total positive circuit conductance just slightly larger than the negative conductance of the tunnel diode, at the operating frequency. The closer the operation is to the resistive cut-off frequency (f_{ro}), the lower the negative resistance becomes since

$$Rd = \frac{1}{gd\left(1 + \frac{\omega^2 c^2}{gd^2}\right)}$$

It is advantageous to operate with relatively large values of negative resistance in order to match conventionally available low loss transmission lines.

b. Stability:

Several "Stability" criteria must be considered:

1. A. C. Stability: This criteria merely states that the circuit is "tamed" into amplifier operation. The circuit cut-off frequency (not to be confused with the device f_{ro}) is made smaller than the circuit self resonant frequency (not to be confused with device f_{xo}) be the addition of load resistance such that $L(\omega) < R_T(\omega) \left|Rd\right| C_T$.

2. D. C. Bias Stability: Since the negative resistance, the cut-off frequency, the tunnel diode capacitance and even the self resonant frequency is bias dependant it is very important for gain, noise and stability considerations, to keep the operating point stable.

3. Coupling Stability: Since gain and stability also depend on matching conditions one must try to minimize the effect of source and load impedance changes. There the Author's have used either Isolators, Circulators or Hybrid Couplers (using two matched tunnel diodes in this latter case), to insure adequate stability.

c. Noise:

Low noise is obtained in "high gain" circuits operating well below the cut-off frequency (f_{ro}). Smoothing effect may make equations look somewhat pessimistic as full shot noise may not be present.

Tunnel Diode Specifications

Resistive cut-off frequency (f_{ro}): as high as possible. Since the negative resistance should also be relatively high, this then essentially means a low current device having very low capacitance and as low as possible series resistance (Rs).

Self Resonant frequency (f_{xo}): as high as possible. Since gd (or Rd) and capacitance (C) have already been commented on, the lowest possible package inductance will yield the highest f_{xo}. This inductance should also be lowest from a stability standpoint.

Series Resistance (Rs): as low as possible. In regards to noise, the ratio of $\frac{Rs}{Rd}$ should be minimized. Rs is frequency dependent due to skin effect.

I_o $|Rd|$ or (Io/|gd|): as low as possible. For minimum noise Io/gd should be minimized. First the tunnel diode manufacturer will have to specify the operating point where Io/gd is at a minimum and its value at that point and then the circuit designer will have to operate there for minimum noise performance. Presently Commercial Germanium units have the lowest Io/gd ratio. Silicon and Gallium Arsenide, due to their greater bandgap have higher Io/gd ratios.

The performance obtained by the Author's of the following amplifier articles are tabulated as follows: (See Page #8)

Author	Operating Frequency	Power Gain	N. F.	$\sqrt{G \times B}$	Saturation Power	Stability	Diode
Yariv etal	4. 5 Kmc	23 db	7 db	275 Mc	1 μω	-	1 ma-Ge
Trambarulo	6. 8 Kmc	38 db	7. 5 db	-	-	-	0. 5-1. 0 ma Ga As.
	10. 8 Kmc	38 db	11. 5 db	-	See Fig. 1	-	
	25. 8 Kmc	36 db	-	-	-	-	
Sie #	210-625 Mc	8. 2 db	1. 93 db	330%	-20 dbm (22mv)	Equiv. Input VSWR < 2	1 ma-Ge
Schaffner	450 Mc	15 db	4. 2-5. 5 db	100 Mc	-	-	1 ma-Ge

In Mr. Sie's paper stability with input VSWR variation is given as: > 100:1 (load VSWR = 1. 05).

(1) "A NOISE INVESTIGATION OF TUNNEL DIODE MICROWAVE AMPLIFIERS"

by

A Yariv and J. S. Cook
Proc. IRE April 1961, pp. 739-743

This article presents an analysis and derivation of the noise figure of a tunnel diode microwave amplifier. The agreement between the measured noise figure and the theoretical results is an indirect check on the existence of full shot noise in germanium tunnel diodes at microwave frequencies. The limiting noise temperature of the amplifier is eIoR/2k, and can be approached by using diodes with small (RC) products in which the extreme overcoupling (load mis-match) and high gain can be achieved simultaneously.

In their introduction Mr. Yariv et al states: "Amplifiers utilizing tunnel diodes were first reported by Chang (1) at frequencies up to 80 mc/s.

He also suggested that the noise contribution of the diode was due to the shot noise accompanying the tunneling process. This suggestion was confirmed by Lee and Montgomery (2) who showed by measurements at 5 mc/s that full shot noise was indeed generated by the tunnel diode. Tieman (3) has reported results similar to those of Lee and Montgomery, obtained from noise measurements at 500 kc on germanium diodes.

Approximate noise analyses of tunnel diode amplifiers were given by Sommers, et al (4), Tieman (3) and Hines and Anderson (5). The simplified equivalent circuits used were sufficient to derive the limiting noise behavior for very high gain and extreme overcoupling. Yariv et al, (6) reported the operation of a tunnel diode amplifier at microwave frequencies and measured its noise figure.

In this paper, the details are given of the noise analysis which led to the noise temperature formula stated by Yariv, et al (6). This analysis is based on an "exact" equivalent circuit, and its results are useful in the intermediate range of small or

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(1) K. K. N. Chang "Low Noise Tunnel Diode Amplifier" Proc. IRE (Correspondence) vol. 47, pp. 1268-1269, July 1959.

(2) C. A Lee and H. C. Montgomery, "Determination of Forward and Reverse Tunneling Currents in Esaki Diodes by Shot Noise Measurement", Bull. Am. Physics Soc., vol. 3, p. 160, March 21, 1960.

medium gain as well as in the limiting case described above. The theoretical results are compared with the noise measurements.

The amplifier Yariv et al constructed consists of a germanium diode mounted in a single port cavity which is coupled to the load and the source by a low-loss circulator as shown in Figure #1, the coupling being continuously variable. The equivalent circuit of this amplifier is shown in Figure #2. The diode is represented by a negative resistance -R, shunted by a capacitance. The spreading and contact resistance of the diode are lumped together with the ohmic losses of the cavity in Rc.

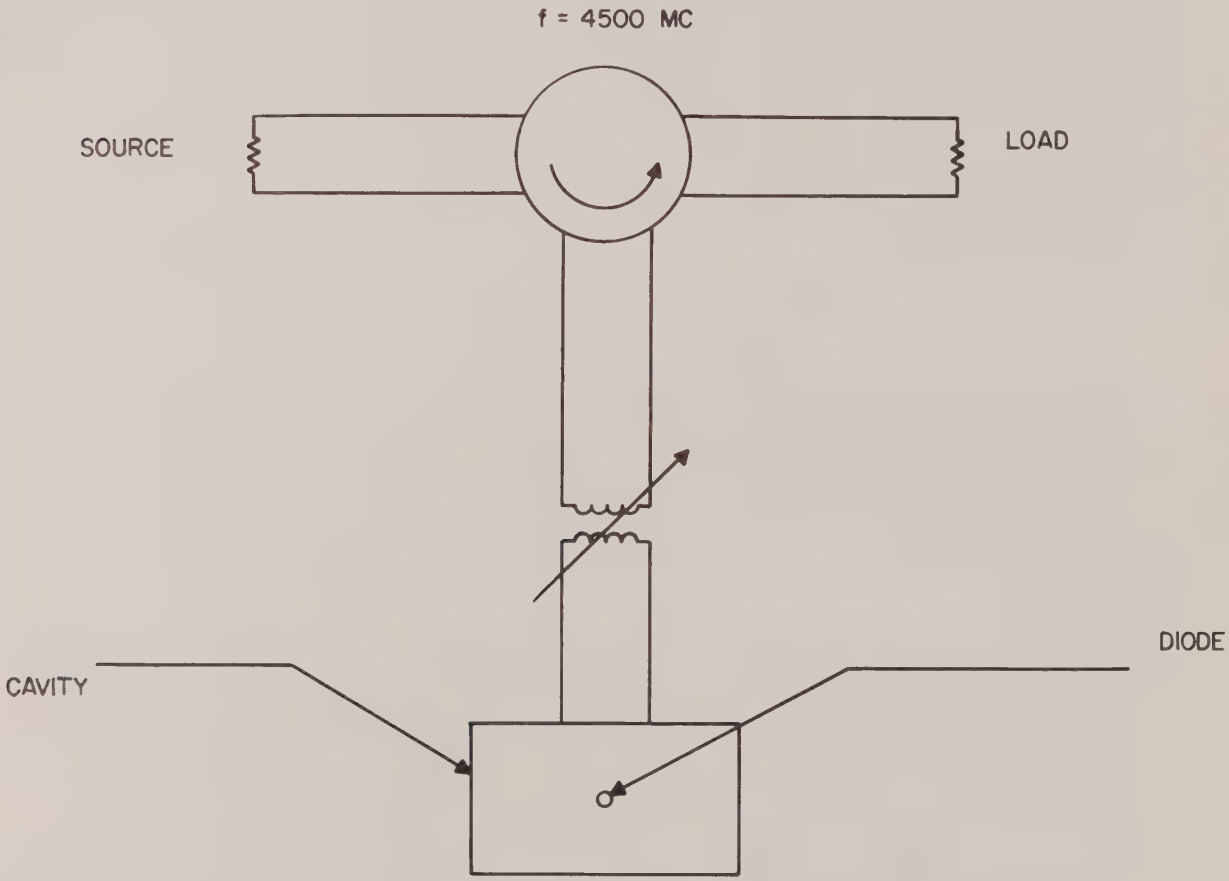

Figure #1 - A schematic view of the Esaki Diode Amplifier

(3) J. J. Tieman, "Shot Noise in Tunnel Diode Amplifiers", Proc. IRE vol. 48, pp. 1418-1423, August 1960.

(4) H. S. Sommers, Jr. et al "Tunnel Diodes for Low Noise Amplification", 1959, IRE Wescon Convention Record, pt. 3, pp. 3-8.

(5) M. E. Hines and W. W. Anderson, "Noise Performance Theory of Esaki (Tunnel) Diode Amplifiers", Proc. IRE (Correspondence) vol. 48, p. 789, April 1960

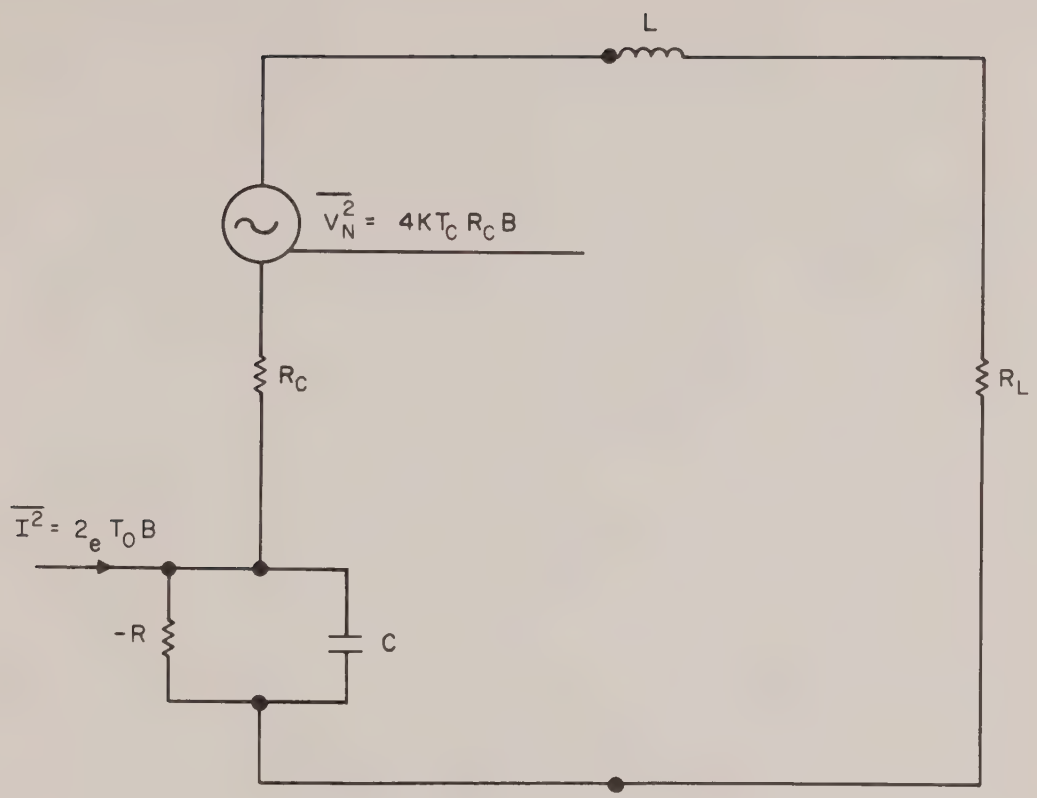

Figure #2 - The Equivalent Circuit of the Esaki Diode Amplifier

For the purpose of analysis, Yariv et al transform the circuit to one in which all the elements are either in series or in parallel. The series equivalent circuit and the new expression for the equivalent series circuit elements are shown in Figure #3.

Series resonance occurs at

$$\omega_{res} = \omega_o \sqrt{1 - \frac{1}{\omega_o^2 R^2 C^2}} \qquad (1)$$

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(6) A. Yariv et al, "Operation of an Esaki diode microwave amplifier", Proc. IRE (correspondence), vol. 48, p. 1155, June 1960.

where $\omega_o^2 = 1/L\,C$. For resonance to occur at real frequencies, one must fulfill the condition

$$\omega_o\, R\, C > 1 \tag{2}$$

for $\omega_o^2\, R^2\, C^2 \gg 1$, $\omega_{res} \approx \omega_o$ (3)

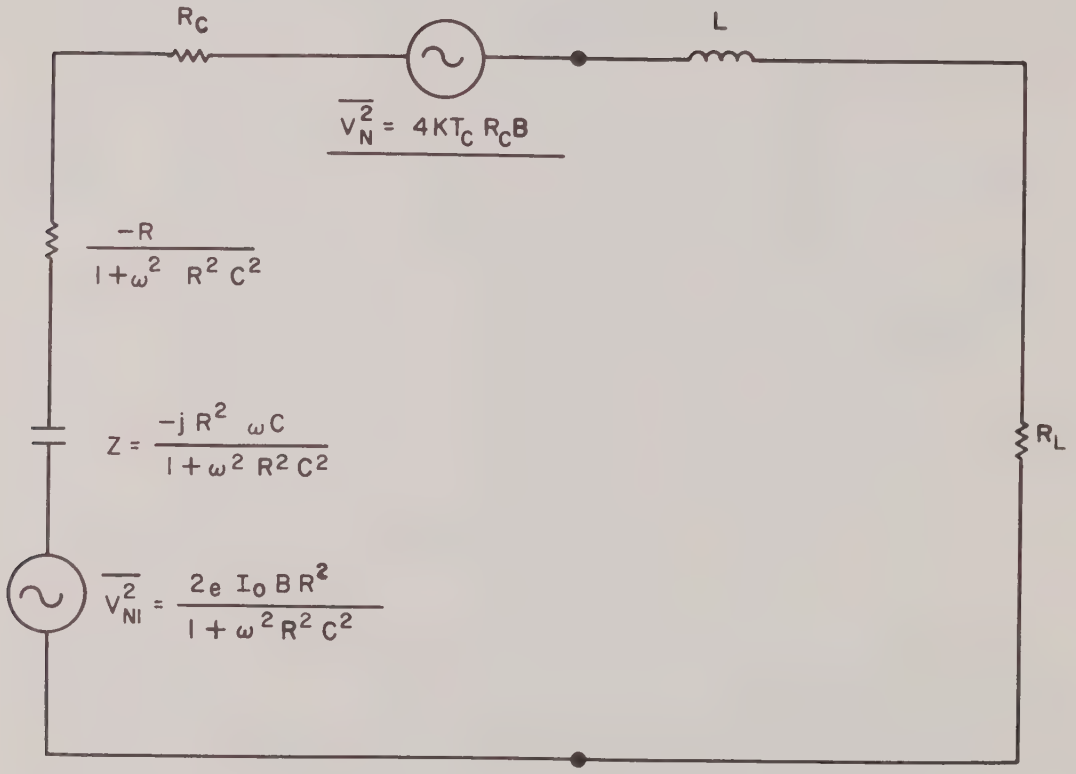


Figure #3 - The Series Equivalent Circuit of the Esaki Diode Amplifier

A necessary and sufficient condition for positive gain is that the real part of the total series resistance, seen looking back into the amplifier terminals, be negative or from Figure #3,

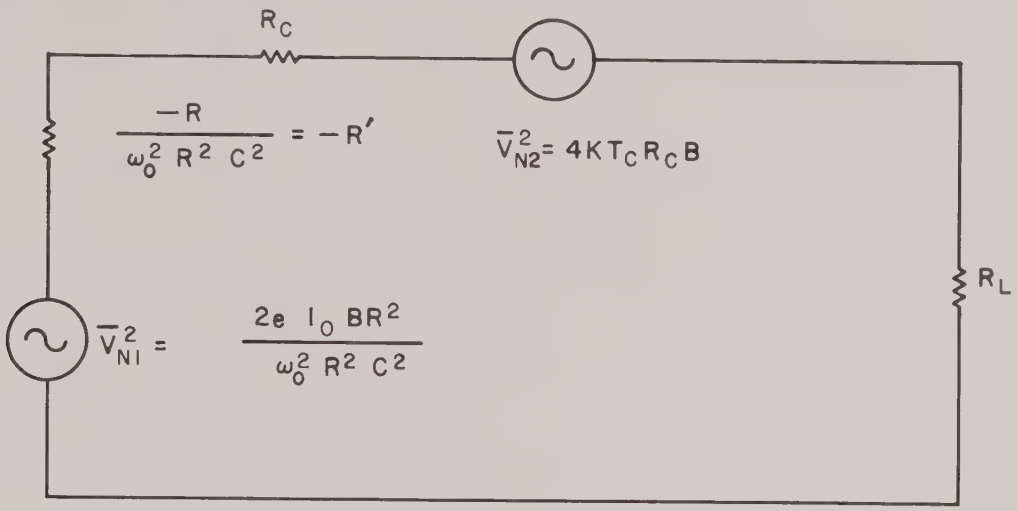
$$\frac{R}{1+\omega^2\, R^2\, C^2} > Rc \tag{4}$$

The highest frequency at which this can occur is called the cutoff frequency, f_{co}:

$$f_{co} = \frac{\sqrt{R/Rc - 1}}{2\pi\, RC} \tag{5}$$

and represents the highest useful frequency for a given diode and circuit.

Mr. Yariv et al. subsequently evaluates the elements of the equivalent circuit in Figure #3, for the condition $\omega = \omega_{res}$. The result is shown in Figure #4.



THE SERIES EQUIVALENT CIRCUIT
OF THE ESAKI DIODE AMPLIFIER
AT RESONANCE

FIGURE 4

The effective negative resistance is now represented by the series resistance.

$$R^1 = \frac{-R}{\omega_o^2 R^2 C^2} \qquad (6)$$

while the shot noise is now provided by a voltage generator of mean-square amplitude.

$$V_{NI}^2 = \frac{2e\, Io\, BR^2}{\omega_o^2 R^2 C^2} \qquad (7)$$

The author points out that physical significance should "only" be attached to ratios of any pair of resistances in Figure #4. The quantity R^1/FL is equal to the ratio of power generated by the diode to that delivered to the load, while RL/Rc, for instance, gives the ratio of the power consumed by the load to that expended in the cavity. RL/Rc is often referred to as the coupling ratio of the cavity. (Ratio of cavity "Q" to

external "Q").

The author further gives the gain of the amplifier as

$$G = \left[\frac{R_L - (Rc - R^1)}{R_L + (Rc - R^1)} \right]^2 \tag{8}$$

and the bandwidth as

$$\Delta f = \frac{R_L + Rc + R^1}{2\pi L \left(1 - \frac{1}{\omega_o^2 R^2 C^2}\right)} \tag{9}$$

yielding for the voltage gain-bandwidth product

$$(\sqrt{G}) \text{ res. } \Delta f = \sqrt{G}\ \Delta f = \frac{R_L + R^1 - Rc}{2\pi L \left(1 - \frac{1}{\omega_o^2 R^2 C^2}\right)} \tag{10}$$

Eq. (10) takes a meaningful form upon substitution of the high gain condition

$$R^1 \approx Rc + R_L$$

(if $R^1 = Rc + RL$, gain becomes infinite, the "start oscillation" condition), and the definition of R^1 as given by (6). The result is

$$(\sqrt{G}\ \Delta f) \text{ high gain} = \frac{1 - Rc/R^1}{\pi RC \left(1 - \frac{1}{\omega_o^2 R^2 C^2}\right)} \tag{11}$$

in which form it is amenable to experimental verification.

To increase the gain-bandwidth product, we have to use a diode with a small (RC). This happens, not only through the dependance of $\sqrt{G}\ \Delta f$ on (RC) as given by (11), but also through a less obvious decrease of Rc/R^1. If we assume that the total inductance has been made as small as possible, operation at one frequency means using the same C. If one now goes to a diode material with an intrinsically smaller (RC) product, it becomes possible to operate at the same frequency with a diode

having a smaller R, i. e. a diode generating more power, which corresponds to a smaller Rc/R^1 ratio. (This new diode junction will have the same area as the old one, but will possess a larger current density and, subsequently, a larger current).

The author then gives the equation for noise temperature of the cavity and diode (Te). (Noise figure (F) is related to Te by $F = 1 + \frac{Te}{290}$).

$$Te = \frac{(\sqrt{G} + 1)^2}{G} \left[Tc\left(\frac{Rc}{R_L}\right) + \frac{eIoR}{2k}\left(\frac{R^1}{R_L}\right)\right] \qquad (12)$$

This form is particularly useful, since in it the dependence of the noise temperature on the gain is separated from its dependance on the coupling ratios.

In the experimental results a 1 ma (Ip) unit with a IoR product of 8.6×10^{-2} at the biasing point, a R of 175 Ω and a RC product of 3.5×10^{-10} $\pm$ 10% was used.

The measured gain-bandwidth product at 4. 5 K Mc was 3.6×10^8, checked at a number of gain settings between 17 and 25 db, where it was found to be a constant. It deteriorated, however, for lower gain settings.

The measured noise temperature was 1740°K taken at 25 db gain. The calculated value for Te was found to be 1890°K $\pm$ 10%. The main source of error is the uncertitude in (RC). The author further states that a more recent version of the same amplifier yielded a noise temperature of 1200°K.

In the summary it is stated that:

a. Te depends on gain so that high gain is imperative.

b. Noise can be minimized by biasing at the minimum IoR point.

c. The IoR product is inherently smaller for semiconductors with small energy gaps (InSb = 0. 17 v, InAs = 0. 33 v, Ge = 0. 72 v) which should therefore be preferred for low noise applications.
   Comment: It should be noted that the bandgap of silicon is 1. 09 v and Gallium Arsenide is 1. 43 v.

d. A small value of RC makes R smaller, at a given frequency, and entails using a microwave structure with low characteristic

impedances. In the diode used in this experiment, $R \approx 175\ \Omega$ and $RC \approx 3.5 \times 10^{-10}$ ($C \approx 2$ pf). Had a RC product of $\approx 10^{-11}$ been used (for the same f, hence C) R would have to have been $\approx 5\Omega$, which is a very low impedance level in microwave circuitry and creates a gamut of problems, such as matching to standard components with much higher impedances and the high attenuation of low-impedance waveguides.

e. For stable operation $L(\omega) < Rs(\omega) RC$ at all frequencies.

The authors then conclude:

"It is not likely that tunnel diode will ever threaten the maser and parametric amplifiers as a low-noise amplifier. It may still find application in cases where its extreme simplicity and economy may be traded for increased noise, especially when tunnel diode amplifiers with noise figures substantially lower than the one reported above will be made. This improvement is mainly contingent on the utilization and taming of diodes with small (RC) products and on the use of low-energy-gap diodes with small IoR products".

Comments on the Article of A. Yariv et al:

This paper describes a 4.5 K mc tunnel diode amplifier. It formulates expressions of gain, bandwidth, gain-bandwidth product as well as noise temperature and hence noise figure for the amplifier. It substantiates, indirectly, the presence of full shot noise at these frequencies by reasonable correlation of calculated versus measured results.

While doing all this the authors have tried to be as practical as possible by considering stability and circuit - as well as device limitations. As a result the authors have come up with practical recommendations of a stable low-noise microwave tunnel diode to be used in microwave amplifiers. Such a unit would have a small IoR product, they would be made of lower bandgap material (presently germanium units are the lowest-commercially-available bandgap devices, Gallium Arsenide the highest with Silicon in between. Another conclusion one can reach from this article is that the diode capacity should be as small as possible.

Finally the authors recommend for practical reasons the use of a relatively large

negative resistance in order to avoid matching problems and the greater loss of low impedance components.

No specific recommendations were made for the values of the series inductance and resistance of the diode package.

(2) "OPERATION OF AN ESAKI DIODE MICROWAVE AMPLIFIER"

by

A. Yariv, J. S. Cook and P. E. Butzien
IRE Proc. (correspondence) June 1960, p. 1155

In this article Mr. Yariv et al give the results of a 4.5 K mc/s tunnel diode amplifier:

Maximum stable gain = 23 db

Voltage Gain-Bandwidth product = 2.75×10^8
(23 db gain at 20 mc/s bandwidth)

Effective source noise temperature = $1200^\circ K$
(noise figure = $\left(1 + \frac{Te}{290^\circ K}\right) \approx 7$ db)

Saturation power output $\approx 1 \times 10^{-6}$ watts

The diode peak current was 1.1 ma and the peak-to-valley ratio was 2.5.

Since this amplifier is thoroughly covered in the previous paper no further comments have been made.

(3) "NOISE PERFORMANCE OF TUNNEL DIODES"

by

E. G. Nielsen
IRE Proc. (correspondence) November 1960
pp. 1903-1904

In his introduction E. G Nielsen points out that it is well known that the negative resistance element of a tunnel diode exhibits shot noise. (1) A tunnel diode also has a parasitic series resistance. If it is assumed that this exhibits thermal noise, some significant conclusions can be drawn. The relative contribution of these two noise sources to the diode's noise performance is frequency dependant. At frequencies much lower than the cutoff frequency, the shot noise is normally the main contributor. At frequencies approaching the cutoff frequency the thermal noise becomes relatively more important and the tunnel diode's noise performance deteriorates.

This performance is determined from the noise equivalent circuit in Figure #1. Here R_D is the diode negative resistance, Rs is the parasitic series resistance, and id and es are, respectively, the shot and thermal noise generators where:

$$\overline{id^2} = 2\ e\ Io\ \Delta f,$$

$$\overline{es^2} = 4\ k\ TRs\ \Delta f$$

and Io is the D. C. current.

For mathematical convenience the shot noise is represented by the equivalent noise conductance Ge where

$$Ge = \frac{eIo}{2\,kt}$$

$$\overline{id^2} = 4\ k\ T\ Gd\ \left(\frac{Ge}{Gd}\right)\ \Delta f$$

$$GD = \frac{1}{R_D}$$

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(1) J. J. Tieman, "Tunnel Diodes and Their Use as Multifunction Circuit Elements" presented at 1960 Internat. Solid State Circuits Conference, Philadelphia, Pa., Feb. 10th, 1960.

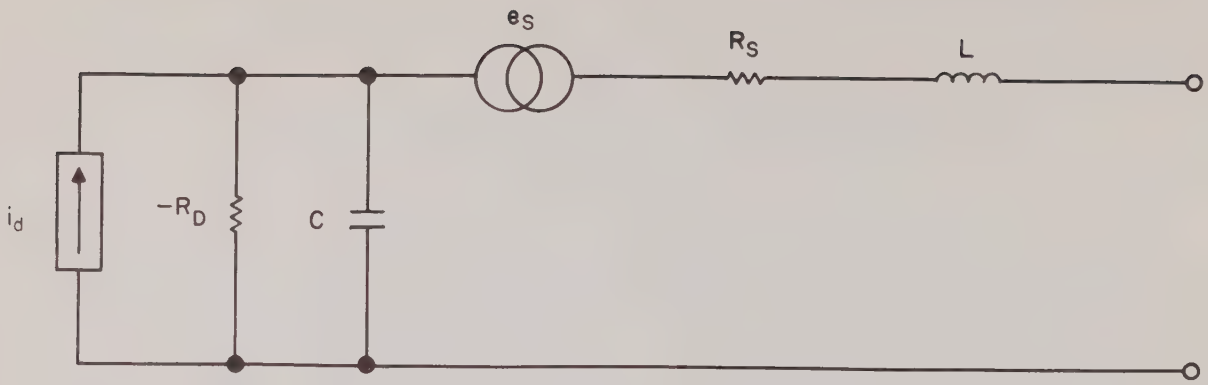

Figure #1 - Noise equivalent circuit

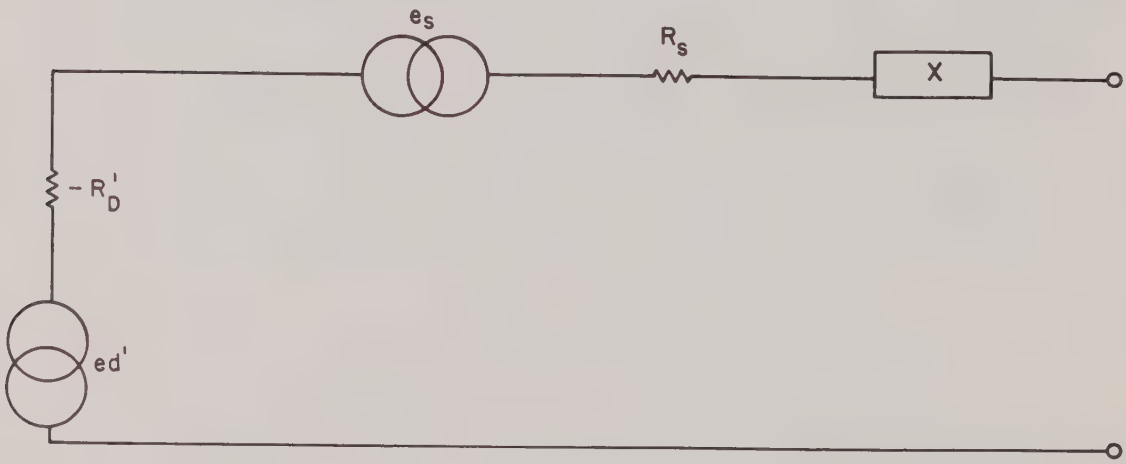

Figure #2 - Single frequency series noise equivalent.

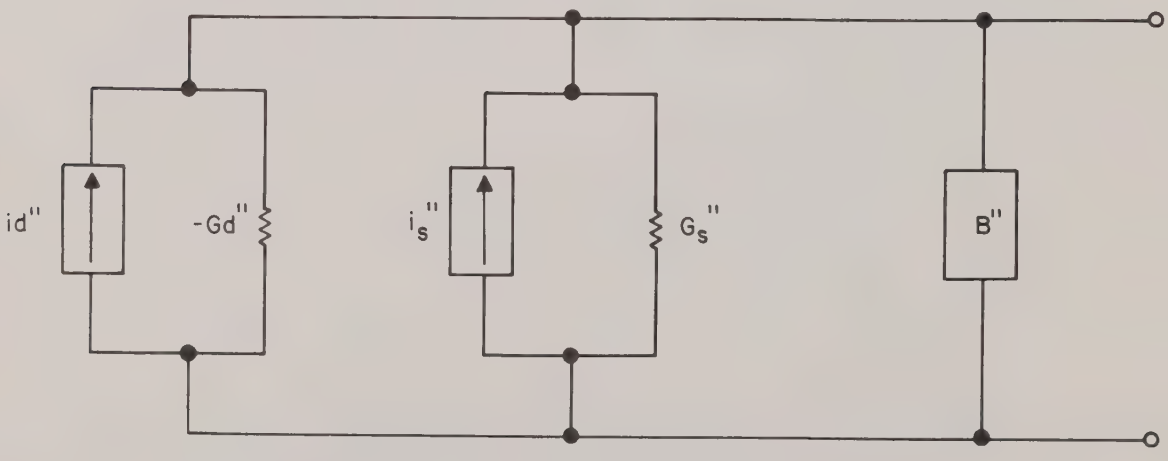

Figure #3 - Single frequency shunt noise equivalent.

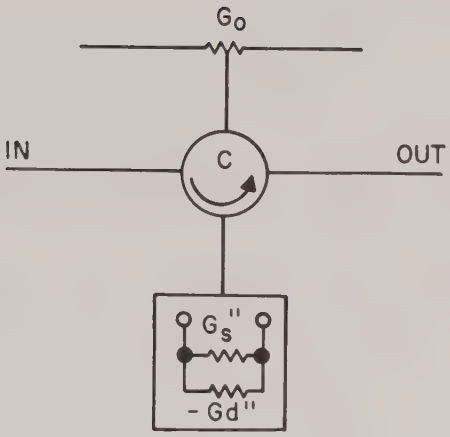

Figure #4 - Tunnel Diode amplifier employing circulator

For calculation of spot noise figure the circuit is put in form of a pure series or shunt equivalent, figures 2 & 3, respectively. In Figure #2, the negative real part of the impedance is reduced by the shunt capacity to the value:

$$Rd^{1} = \frac{Rd}{1 + \omega^{2} Rd^{2} C^{2}}$$

the reactance is given by:

$$jX = j\omega \left[L - \frac{C\, Rd^{2}}{1 + \omega^{2} Rd^{2} C^{2}} \right]$$

Mr. Nielsen subsequently states: "A negative resistance amplifier can be made unilateral by means of a circulator, Figure #4. This arrangement also achieves the lowest possible noise measure. (4) The gain becomes high when the negative admittance approaches the characteristic admittance of the circulator. Assuming the signal source admittance Gg and circulator characteristic admittance Go are real and equal, and the tunnel diode reactance B" is tuned out, this gives

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(4) H. A. Haus & R. B. Adler, "Circuit Theory of Noisy Networks", John Wiley & Sons, Inc. New York, New York, pp. 68-71; 1959.

$$Gg = G_o = G_d'' - G_s'' + \epsilon$$

where ϵ is an arbitrary small positive number. In the limit, as ϵ approaches zero, the gain approaches infinity and the noise figure becomes (5)

$$F = 1 + \frac{T\left(\frac{Ge}{Gd} + \frac{Gs''}{Gd''}\right)}{To\left(1 - \frac{Gs''}{Gd''}\right)} \qquad (6)$$

where To is a reference temperature. Equation (6) is also significant to two port tunnel diode preamplifiers. It can be shown that (6) is the minimum noise figure obtainable with such a preamplifier when it is followed by additional relatively noisy circuits. If it is assumed that T = To (6) becomes a function only of the tunnel diode. It then is a very useful index of the diode's noise performance.

Assuming T = To and making use of (2) - (5),

$$F = 1 + \frac{\left(1 + \frac{Ge}{Gd}\right)}{\left(1 - \frac{Rs}{R_D}\right)\left[1 - \left(\frac{f}{fc}\right)^2\right]} \qquad (7)$$

where fc is the cutoff frequency

$$fc = \frac{1}{2\pi CR_D}\sqrt{\frac{R_D}{R_S} - 1} \qquad (8)$$

The author concludes:

"The noise figure may be considered a function of three ratios, Ge/Gd, Rs/R_D, and f/f_c. The first assesses the shot noise relative to the negative conductance, the second, the parasitic series resistance relative to the negative resistance; and the third, the operating frequency relative to the diode's cutoff frequency. With germanium tunnel diodes $Rs/R_D \approx 0.02$ and $Ge/Gd \approx 1$. Thus at frequencies much smaller than fc, a noise figure of 3.1 db is obtained; and this is almost entirely due to the diode shot noise. As the operating frequency approaches fc the noise figure deteriorates rapidly. For example, the increase is 3 db for f = .707 fc, and

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(5) M. E. Hines & W. W. Anderson, "Noise Performance Theory of Esaki (Tunnel Diode Amplifiers", Proc. IRE, Vol. 48, p. 789, April 1960.

7.2 db for f = 0.9 fc. The frequency function is very important in microwave amplifiers.

Comments on Mr. Nielsen's paper:

Mr. Nielsen points out in his paper that at microwave frequencies the diode parasitic series resistance becomes a very important contributor to the total noise figure. He derives an equation that permits an appreciation of this noise contribution. His equation should be very helpful in the choice of the best diode to be used for operation at a given frequency. Namely the highest cutoff frequency unit, having the most advantageous R_S/R_D ratio. The expression in the numerator is minimized by operating at the minimum Io/Gd point.

(4) "THEORETICAL JUSTIFICATION FOR SHOT-NOISE SMOOTHING IN THE ESAKI DIODE"

by

R. La Rosa and C. R. Wilhelmsen
IRE Proc. (correspondence) November 1960, p. 1903

In their article La Rosa and Wilhelmsen "advance a crude theory" of shot noise smoothing in the negative conductance region. They state that the theory described predicts that most of the eligible electrons do not have infinitesimal probability of tunneling and therefore, the diode need not be as noisy as first assumed. According to the full shot noise theory,

$$\overline{i^2} = 2e\,\Delta f\,(I_{c\rightarrow v} + I_{v\rightarrow c}), \qquad (1)$$

where e is the electronic charge, Δf is the width of a rectangular shaped incremental passband, $I_{c\rightarrow v}$ is the long time average current caused by n-region conduction electrons tunneling into p-region valence states, and $I_{v\rightarrow c}$ is the long time average current caused by p-region valence electrons tunneling into n-region conduction states. The net direct current is,

$$I = I_{c\rightarrow v} - I_{v\rightarrow c} \qquad (2)$$

The authors then say that $I_{v\rightarrow c}$ is not smoothed because it comes from a small probability mechanism. Its shot noise should agree with the second component of (1).

The first component of (1) is reduced by smoothing, and the total mean-square noise current should be,

$$\overline{i^2} = 2e\,\Delta f\,(\gamma^2 I_{c\rightarrow v} + I_{v\rightarrow c}) \qquad (3)$$

where γ^2 is a smoothing factor roughly equal to that assumed by Hines[4]. The author then attempts to find a rough value for γ^2. An approximate "rough" value deduced is 0.5, hence the total shot noise might be reduced considerably.

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4 Hines, M. E., "High Frequency Negative Resistance Circuit Principles for Esaki Diode Applications", Bell System Technical Journal, Vol. 39, pp. 477-513, May 1960.

The authors conclude that not all tunnel diodes should show this noise reduction. It might be significant that the experiments of Chang, et al[8], use diodes having fairly large peak currents, indicating a large ratio of junction area to periphery. This could reduce the surface complications and improve the smoothing.

Comments on Mr. LaRosa's and Wilhelmsen's Paper

The theory advanced by the author indicates hope for lower than expected shot noise contribution of tunnel diodes. It should be an area of emphasis to better understand and make use of this principle in the manufacture and application of tunnel diodes.

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8 Chang, K. K. N, et al, "Low Noise Tunnel Diode Down Converter Having Conversion Gain", Proc. I. R. E., Vol. 48, pp. 854-858, May 1960.

(5) "ESAKI DIODE AMPLIFIERS AT 7, 11, AND 26 KMC"

by

R. F. Trambarulo
IRE Proc. (correspondence December 1960, pp. 2022-2023

The author has operated gallium arsenide Esaki diode amplifiers with stable gains up to 38 db at 6.8 and 10.8 Kmc and 36 db at 25.8 Kmc. The basic amplifier consists of a reflection cavity containing a 0.5 - 1 ma GaAs diode with the input and output separated by means of a circulator or directional coupler.

The cavity is cylindrical in shape[1] with the diode shunting the apex of a re-entrant cone to the opposite face. The cavity operates in the dominant mode for which the diameter is essentially half of a wavelength. Microwave power is coupled in and out of the cavity by means of a large coupling loop on the end of a coaxial line fed from a waveguide. Except at the highest frequency, a small range of loop positions within the cavity could be found over which oscillation could be quenched in the entire negative resistance region of the diode characteristic. With the loop set in the center of this range, a slide-screw tuner in the waveguide served as a fine control on the gain which could be varied from 5 to 38 db by insertion of the screw. At 25.8 Kmc the loop was soldered to the interior of the cavity, and all tuning adjustments had to be made in the waveguide.

The author then says that when the gain is high, it becomes particularly sensitive to small changes in the negative resistance (or average negative resistance for finite amplitude signals). At the bias for maximum gain the diode resistance is a minimum, and therefore the gain drops for small bias changes or for finite amplitude signal swings. Figure 1 shows the Saturation Characteristic of a 10.8 Kmc amplifier.

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1 Trambarulo, R, and Burrus, C. A., "Esaki Diode Microwave Oscillators", Proc. I. R. E., Vol. 48, p. 1176, October 1960.

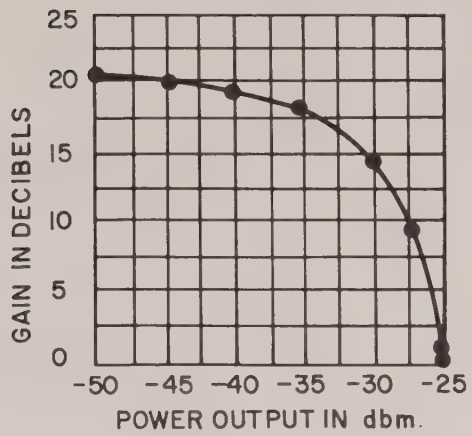

Figure #1 - Saturation Characteristic of a 10.8 Kmc Amplifier

Figure #2 shows the bandwidth versus gain for a 10.8 Kmc amplifier. As is typical of a reflection amplifier, the gain and frequency response were found to be dependant upon reflections from the circulator. With a poorly matched circulator, peaks in gain could be found at several frequencies and a strong tendency to break into oscillation often interfered with gain adjustments.

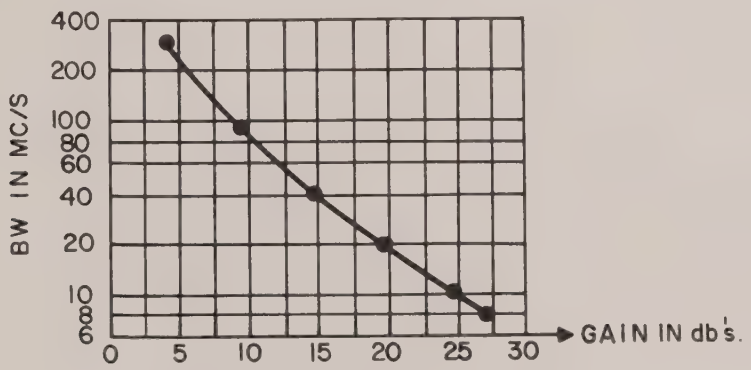

Figure #2 - Bandwidth Gain Relationships for 10.8 Kmc Amplifier

Measured noise figures were 7.5 db at 6.8 Kmc (25 db gain), 10.7 db at 9.7 Kmc/s (22 db gain) and 11.5 db at 10.8 Kmc (26 db gain). No noise measurements were made at 25.8 Kmc.

Comments on Mr. Trambarulo's Paper

This paper shows interesting performance characteristics at very high frequencies. Bias stability problems are brought to the readers attention (but could probably be overcome by bias stabilization schemes)[2].

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2 Tunnel Diode Manual, General Electric Company, 1961, Chapter 3, pp. 26-29.

(6) "ABSOLUTELY STABLE HYBRID COUPLED TUNNEL DIODE AMPLIFIER"

by

John J. Sie

IRE Proc. (correspondence) July 1960, p. 1321

In his work Mr. Sie has used two matched germanium tunnel diodes of the 1N2939 type and a quarter-wave stripline coupled hybrid (Sage 750) in a wide band (210-625 Mc) U. H. F. amplifier circuit. His preliminary tests showed reasonable gain, extremely wide bandwidth, low noise, and the circuit is both open and short circuit stable without any non-reciprocal device.

The block diagram is shown in Figure #1. By proper wave cancellation, the hybrid coupled amplifier is a reciprocal bilateral amplifier with matched positive input and output impedance under ideal circumstances. [1, 2]

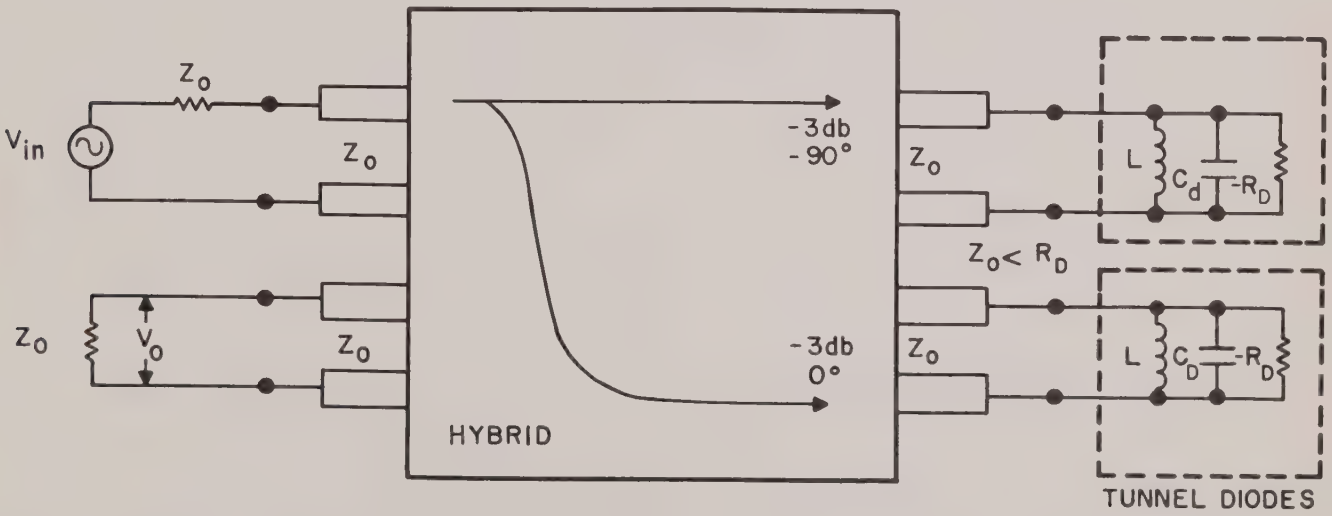

Figure #1 - Block Diagram of Hybrid Coupled Tunnel Diode Amplifier

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1 Sie, J, and Weisbaum, S, "Noise Figure of Receiver Systems using Parametric Amplifiers", 1959 I. R. E. National Convention Record, pt. 3, pp. 141-157

2 Antler, S. H, "Proposed for a Maser Amplifier Without Non-reciprocal Elements", Proc. I. R E., Vol. 46, pp. 1880-1881, November 1958.

These conditions are:

1) Ideal Hybrid Action
2) Matched Negative Resistances
3) Matched Source and Load Impedances

Two General Electric germanium tunnel diodes (1N2939) are used. The diodes have a cut-off frequency of 2.1 Kmc. They were mounted on type-N connectors.

The experimental data is shown in Figure #2.

Frequency Range:	210-625 Mc
Gain:	8.2 db $\pm$ 0.6 db
Noise Figure at 365 Mc:	
a) Measured:	1.93 db $\pm$ 0.4 db*
b) Calculated:	1.5 db
Equivalent Input VSWR:	< 2.0
Gain 1/2 Percentage Bandwidth:	330%
Dynamic Range	
a) $\pm$ 1 db	78 db
b) $\pm$ 3 db	91 db
c) Input Level at -3 db deviation	-20 dbm (22 mv)
Input-Output Connections:	50Ω Type N
Tunnel Diodes:	G-E 1N2939
Hybrid	Sage 750
Stability with Input VSWR Variation:	> 100:1 (Load VSWR = 1.05)

*Limit of experimental accuracy

Figure #2 - Experimental Results of Hybrid Coupled Tunnel Diode Amplifier

The author concludes with the following statement:

"The above results represent a usable low noise regenerative amplifier with several orders of magnitude performance improvement over existing devices. The same principle can be used to extend this amplifier to higher frequencies as better tunnel diodes become available.

Comments on Mr. Sie's Article

Mr. Sie has attacked the amplifier stability problem by using two matched tunnel diodes with a hybrid coupler. His concluding statement is especially interesting insofar as better tunnel diodes (greater f_{ro} and f_{xo} devices) have recently become available. For specifications of such units, see pages 86 - 87 of this report.

(7) "A COMPACT TUNNEL DIODE AMPLIFIER FOR ULTRA HIGH FREQUENCIES"

by

Dr. G. Schaffner
IRE-WESCON Convention Record, 1960, pp. 86-93

The stated purpose of Dr. Schaffner's paper is to describe the design and performance of a compact tunnel diode amplifier for the 405 - 460 Mc frequency range. In addition to small size and light weight, this amplifier features a low noise figure of 5.5 db, (Dr. Schaffner comments that a later version only had 4.2 db noise figure) and a relative independance of antenna impedance variations when used with a miniature ferrite isolator. Further the overall (gain) $^{1/2}$ times bandwidth product is close to 100 Mc. With the advantage of small size, low noise figure, and excellent stability, the amplifier should prove highly useful for mobile, airborne, missile and satellite receivers.

Design of the amplifier is based on tuning the reactive elements of the tunnel diode with the external circuit. Coupling to the source and load are adjusted so that the equivalent positive conductance exceed by a slight amount the negative conductance of the diode.

A shorted transmission line is used in place of a lumped coil to tune the tunnel diode capacitance. Considering the large capacitance normally associated with the tunnel diode, the necessary transmission line can usually be made quite short. Equation (1) is the significant design equation relating, at a given center frequency, transmission line length and characteristic admittance of the diode capacitance,

$$\omega_o C_d = Y_o \, C_{tn} \, \frac{\omega o l}{c} \qquad (1)$$

where: ω_o = is the desired center frequency

Y_o = is the transmission line charact. admittance

l = is the length of the shorted line

c = is the velocity of light

One troublesome aspect in the design of any tunnel diode amplifier is the

biasing circuit. Since the diode exhibits its negative resistance from DC to microwaves, a spurious low-frequency oscillation in the biasing circuit often occurs. This oscillatory tuned circuit consists of bias lead inductance and shunt capacitance. To prevent these instabilities the net conductance of the bias circuit at the diode terminals must always exceed the diode negative conductance. This means low-impedance bias circuits (including low inductance) are required. The gain equation:

$$K = \frac{4\, G_g\, G_L}{(G_g + G_L - G)^2}$$

and the noise figure equation:

$$F = 1 + \frac{T}{T_o}\left[\frac{G_l}{G_g} + \frac{G_L}{G_g} + \frac{G_e}{G_g}\right]$$

where $G_e = eI_{eq}/2KT$

show that the input and output couplings are highly important to the achievement of both low noise figure and stable operation. The analysis which is applicable to both tunnel diode and parametric amplifiers, shows that for lowest noise figure in a two-port device the input coupling should be greater than the output coupling ($G_g > G_L$). Accordingly in Dr. Schaffner's amplifier, the input port of the amplifier is coupled directly to the center conductor of the cavity at a high impedance point. The output port, on the other hand, is coupled directly to the center conductor, near the short circuited end.

Dr. Schaffner used a ZJ56 (1N2940) in his amplifier. To provide a low series inductance mounting for this TO-18 packaged device, two parallel leads of the tunnel diode are initially soldered onto a 1000 pf button capacitor, which is an inherent part of the cavity and plate. The other side of the tunnel diode, which is connected to the diode housing, is slipped into a clip soldered directly on the cavity center conductor. For this diode mounting, the total lead length between r-f ground and the tunnel diode is less than 0.1 inch.

Tuning of the amplifier over the 405 - 460 Mc range is accomplished by a coaxial capacitor at the high impedance end of the cavity. This tuning capacitance adds to that of the tunnel diode and is significant in reducing the length of the amplifier.

The schematic diagram is shown in Figure 1.

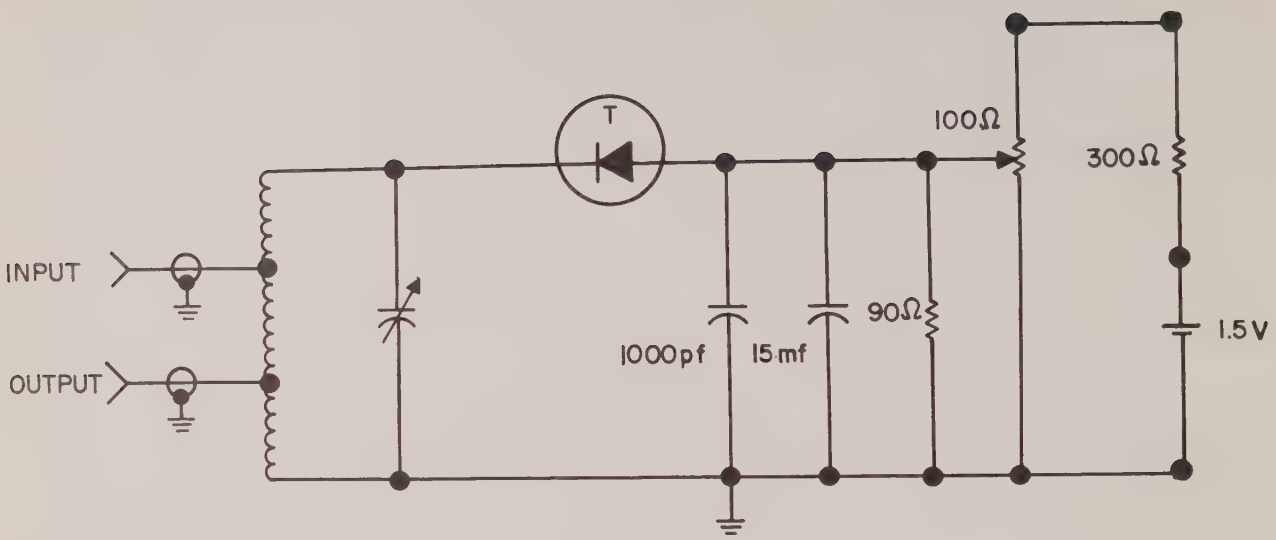

Figure #1 - Schematic of 450 Mc Tunnel Diode Amplifier

Performance of Experimental Amplifier

A gain of approximately 15 db is achieved over the band. The associated bandwidth at 15 db gain varies from 10 Mc at 405 Mc through a maximum of 18 Mc at 420 Mc to 12 Mc at 450 Mc. The associated gain-bandwidth products for a single tuned unit are better than those quoted for parametric amplifiers, probably because of the heavy coupling used and the lack of an idler frequency.

Of most importance, the noise figure obtained from this amplifier is a fairly constant 5.5 db over the band. With the addition of the UHF isolator, which has a forward insertion loss of 0.75 db and an isolation of 12 db at 450 Mc, noise figures of 6.1 db to 6.4 db are obtained. Actually a noise figure of 4.2 db excluding the isolator, has been measured in a later version of the tunnel diode amplifier. With the isolator, system gain changes less than 3 db with antenna VSWR variations from 1.1 - 1.8 db.

Dr. Schaffner concludes that "overall performance indicates that the amplifier is suitable for numerous low noise applications. It could easily reduce the noise figure of an all-transistorized receiver to be comparable to that of a vacuum tube receiver. It should also act as an ideal sensitivity booster in a system having a para-

metric amplifier or maser in the ground or fixed station, with the tunnel diode amplifier installed in the remote receiver."

Comments on Dr. Schaffner's Article

Dr. Schaffner's paper describes the practical design of a UHF amplifier having good gain, (15 db), low noise (4.2 - 5.5 db) and adequate bandwidth (10 - 18 Mc). He emphasizes stability aspects and reduces the sensitivity to input circuit impedance variations by using a small ferrite isolator getting only 3 db change in gain for VSWR $\leq$ 1.8 db.

SECTION II

OSCILLATORS

ABSTRACT:

The papers in this section deal with two distinct classes of operation. In the first two papers the authors have tried to obtain as much power output as possible at a given frequency and had to resort to a method of operation permitting them to operate above "self resonance". Hence they could use higher current (power) units and still operate at very high frequencies.

In the next two papers the authors have generally tried to reach the highest operating frequency and have considered power output secondary. The last paper in this section deals with a specific method of frequency tuning the oscillator, namely by D. C. bias variation.

"Power" Oscillators Performance:

In the first two papers Sterzer et al, and Kim et al. operate above "self resonance" by making the tunnel diode (being inductive above self-resonance) look into a capacitive (or negative inductance) circuit. Kim points out that one problem area is to avoid undesirable oscillation at lower frequency, since the external circuit might look inductive there and the resultant oscillation might become the dominant mode.

Performance obtained can be seen in Table #II of Sterzer's etal. article in this section. One example is a 610 Mc oscillator having 10 mw power output using one 210 ma GaAs diode. Another example shows a 7130 Mc oscillator having 0.012 mw (12 $\mu\omega$) power output using a 13.7 ma Ge tunnel diode.

Highest Frequency Performance:

Trambarulo and Burrus try for the highest frequency performance and prefer GaAs because they feel they can obtain better power as well as gd/c ratios and hence higher power at higher frequencies. The tunnel diode they have used have peak currents in the 0.3 - 2.0 ma region. The highest frequency reported was 103 K Mc (2.9 mm) with a zinc doped GaAs - point contact type unit having

a 600 μa Ip. Their results can be summarized as follows:

Diode	Freq.	Power Output
300-700 μa	39.9 KMc	0.2 μω at 37 K Mc
300-700 μa	17-25 K Mc	3.5 μω
0.5-2.0 ma	16.7 K Mc	50 μω
0.5-2.0 ma	50 K Mc	25 μω
0.5-2.0 ma	90 K Mc	2 μω

Tuning ranges of 10 - 25% typical and 33 - 38% maximum were obtained by piston tuning in waveguides, 0.5 - 2.0% by varying bias.

The authors point out, that the frequency is bias sensitive. In the last paper, Mr. Pulfer deals with just this phenomena insofar as he deliberately uses the bias dependance of the frequency for frequency tuning purposes. He achieves 12% tuning range at 1500 Mc.

(8) "TUNNEL DIODE MICROWAVE OSCILLATORS"

by

F. Sterzer and D. E. Nelson
Proc. IRE, April 1961, pp. 744-753

This article starts with a discussion on the basic tunnel diode parameters and develops into a small signal analysis of the equivalent circuit. Graphs are plotted of the real and imaginary parts of Z_d / R_d as a function of frequency to aid in the design of the oscillators using lumped circuit parameters subdivided into two parts:

1) circuits for operation below the maximum self-resonant frequency of the diode, and

2) circuits for operation above the maximum self-resonant frequency of the diode.

Another section subsequently discusses oscillators using distributed circuits.

In the author's introduction it is stated that, "this paper discusses the use of tunnel diodes in UHF and microwave oscillators with milliwatt power outputs. The diode parameters which determine the performance of the diode in oscillator circuits are described in detail. Several oscillator circuits are analyzed with special attention to the steady state. A number of practical oscillator circuits are described, and experimental results are presented".

An approximate equivalent circuit for an encapsulated tunnel diode consists of three elements connected in series: an inductance L_d, which results mainly from the inductance of the housing, r_d is the resistance of the ohmic contact, the base, and the internal leads of the package, and is a function of frequency due to skin effect, and C_d (v) which is the somewhat voltage sensitive junction capacitance paralleled by $R_d(v)$ which is the total resistance of the junction, where v is the voltage across R_d and C_d.

The dc bias voltage V_d across the tunnel diode is given by:

$$V_D = I_d \, rd + I_d \left[R_d(v) \right] \qquad (1)$$

where I_d is the direct current through the diode. At a reverse current, an order of magnitude greater than I_p, $\left[R_d \right]$ is usually very small compared to rd, so that rd can be measured directly.

In ac application, the quantity of most interest is not the total value of the junction resistance $\left[R_d(v)\right]$, but rather its incremental value $R_d(v) = dv/dI_d$. This value is negative between the peak and valley points of tunnel diodes. The series inductance L_d and junction capacitance $C_d(v)$ can be determined from ac impedance measurements. Theoretically, the variation of C_d with voltage when V_d is less than V_v may be approximated by:

$$C_{d(v)} \approx K (\phi - v)^{-1/2} \tag{2}$$

where K and ϕ are constants.

In terms of the incremental resistance R_d, the small signal ac impedance Z_d across the terminals of a tunnel diode may be written as follows (when diode is biased in the negative conductance region):

$$Z_d = \left(rd - \frac{R_d}{R_d^2 C_d^2 \omega^2 + 1} \right) + j \left(\omega L_d - \frac{R_d^2 C_d \omega}{R_d^2 C_d^2 \omega^2 + 1} \right) \tag{3}$$

In order that the diode exhibit negative resistance, $|R_d|$ must be greater than rd. The real part of the impedance, $R_e(Z_d)$, is equal to $(rd - |R_d|)$ at zero frequency and increases monotonically with frequency. The frequency at which $R_e(Z_d)$ becomes zero is called the cutoff frequency f_c, and is given by:

$$f_c = \frac{\sqrt{|R_d| / rd - 1}}{2\pi |R_d| C_d} \tag{4}$$

At frequencies above f_c, $R_e(Z_d)$ is positive. The imaginary part of Z_d, $I_m(Z_d)$, becomes zero at frequency f_r, where

$$f_r = \frac{1}{2\pi} \sqrt{\frac{1}{L_d C_d} - \frac{1}{R_d^2 C_d^2}} \tag{5}$$

Usually f_r is referred to as the self-resonant frequency of the diode. Below self-resonance the reactance of the diode is capacitive above self-resonance, it is inductive.

Both f_c and f_r are functions of dc bias. If $|R_d| > 2rd$, $|f_c|$ max. occurs at the bias voltage corresponding to $|R_d|$ min. (neglecting the small effect of the

variation of C_d with bias). For $|R_d| < 2rd$, $[f_c]$ max. is at a lower bias voltage. $[f_r]$ max. always occurs at the bias voltage, V_p.

The design of distributed tunnel diode oscillators requires a plot of diode impedance (or admittance) as a function of frequency. In Figure 5, $R_e(Z_d)/|R_d|$ is plotted as a function of $\omega C_d |R_d|$, with $rd/|R_d|$ as a parameter. Figure 6 shows $I_m(Z_d)/|R_d|$ as a function of $\omega C_d |R_d|$, with $L_d/C_d R_d^2$ as a parameter.

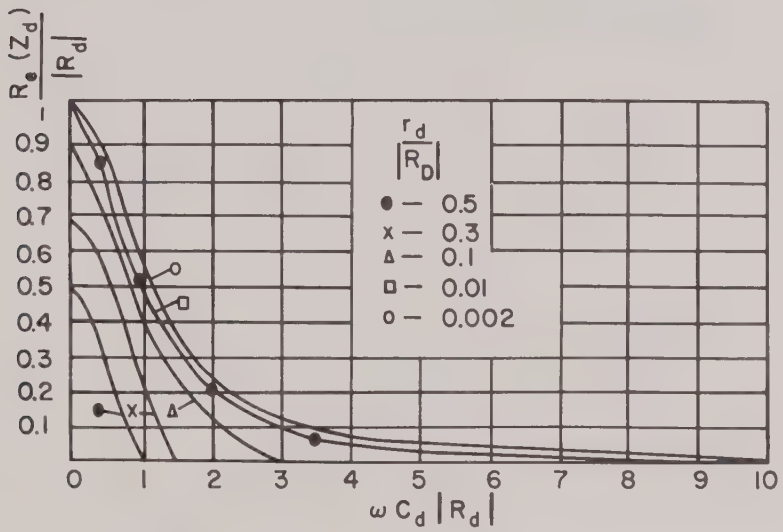

Figure 5 - Graph of the Real Part of $(Z_d)/|R_d|$ as a function of $\omega C_d |R_d|$ with $rd/|R_d|$ as a parameter

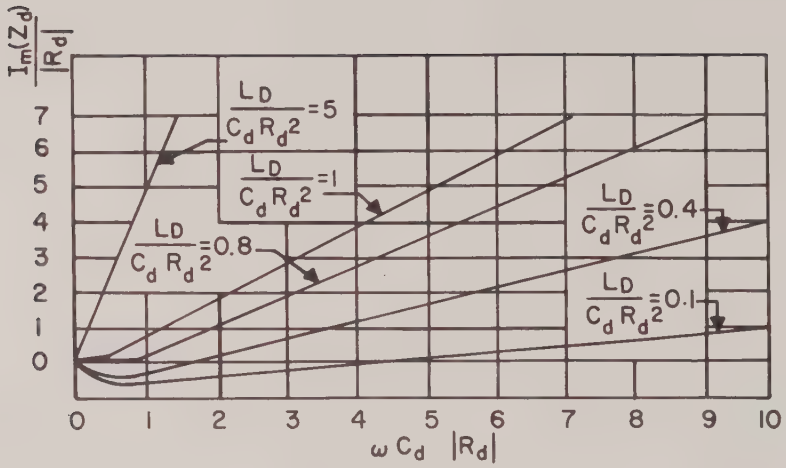

Figure 6 - Graph of the imaginary part of $(Z_d)/|R_d|$ as a function of $\omega C_d |R_d|$ with $L_d/C_d R_d^2$ as a parameter

The authors then state:

From an analytical standpoint, the simplest possible tunnel diode oscillator circuit consists of a diode shunted by the series combination of an inductance L_c, a series resistance r_c, and a dc voltage source V_b, as shown in Figure 7a.

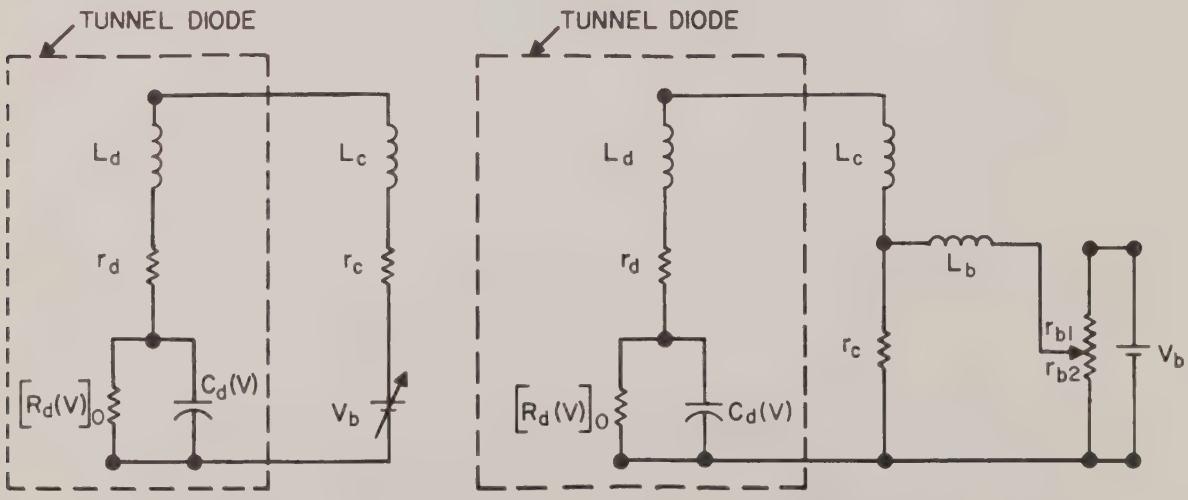

FIGURE 7a FIGURE 7b

Tunnel Diode Oscillator Circuits for operation below the Maximum Self-resonance Frequency of the Diode

In many tunnel diode oscillator circuits, the inductance of the leads from the DC supply is much greater than the desired value of L_c. The problem of lead inductance can be obviated through the use of the circuit shown in Figure 7b. Here the resistor r_c can be placed close to the diode, and the inductance of the bias leads L_b does not affect the performance of the circuit probided both r_{b1} and r_{b2} are much larger than r_c.

The initial response of the network to a small signal excitation can be determined by taking the Laplace transform of equations (6) and (7). The resultant characteristic equation is given by:

$$S^2 L C_{di} + S (rC_{di} - L/ \left|R_{di}\right|) + (1 - r/ \left|R_{di}\right|) = 0 \qquad (8)$$

where R_{di} and C_{di} are the initial values of R_d and C_d, respectively. Equation (8) can be solved for S:

$$S_{1,2} = \sigma 1 \pm j\omega_1 \qquad (9)$$

$$S_{1,2} = -1/2\left(\frac{r}{L} - \frac{1}{C_{di}\ R_{di}}\right) \pm \left[1/4\left(\frac{r}{L} - \frac{1}{C_{di}\left|R_{di}\right|}\right)^2 - \frac{1-r/\left|R_{di}\right|}{L\ C_{di}}\right]^{1/2}$$

a growing solution is obtained if $\sigma_1 > 0$. This condition for σ_1 is satisfied if either one or both of the following inequalities hold:

$$L > r \left|R_{di}\right| C_{di} \qquad (10)$$

$$r > \left|R_{di}\right| \qquad (11)$$

The conditions for sinusoidal growth are inequality (10) and

$$\frac{1}{L\ C_{di}} > \frac{1}{4}\left(\frac{r}{L} + \frac{1}{C_{di}\left|R_{di}\right|}\right)^2$$

The initial frequency of oscillation ω_1 is given by the second term of (9). The steady-state frequency ω_s differs, however, from ω, because of the nonlinearities of the diode.

$$\omega_s \approx \left[\frac{1}{(Ld + Lc)\ Cd} - \frac{1}{Rd^2\ Cd^2}\right]^{1/2} \qquad (19)$$

$$= \left[\frac{1}{(Ld + Lc)\ Cd} - \frac{r^2}{(Ld + Lc)^2}\right]^{1/2}$$

where $\overline{R_d}$ and $\overline{C_d}$ are average values of R_d and C_d and $r = L/\overline{C_d}\,\left|\overline{R_d}\right|$.

Comparison of ω_s (eq. 19) and f_r (eq. 5) shows that ω_s is less than or equal to $\left[\omega_r\right]_{max}$, i. e., the circuits of Figure 7a & 7b can not produce oscillation above the maximum self-resonance frequency of the diode.

The approximate power delivered by the diode is:

$$P \approx \frac{I}{8}\left(V_v - V_p\right)\left(I_p - I_v\right)$$

Above self-resonance, the reactance of the diode is inductive. Therefore, it is necessary to use a circuit presenting a capacitive reactance at the oscillation frequency. A simple circuit suitable for operation above the self-resonance frequency of the diode is shown in Figure 10 (a). However, the circuit shown in 10 (b) must be used if the inductance of the power supply leads L_b in appreciable. The results for the circuit in 10 (a) are also applicable to the circuit in 10 (b), provided r_{b1} and r_{b2} are much greater than r_c.

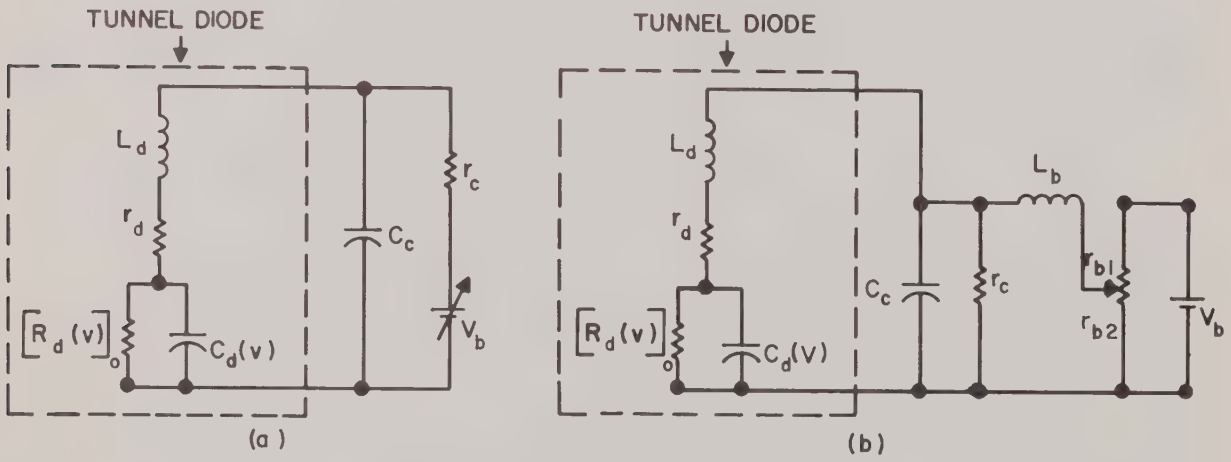

Figure 10 - Tunnel Diode Oscillator Circuits for Operation Above the Maximum Self-resonance Frequency of the Diode

Proceeding in the same fashion as in the preceding section, it can be shown that for small amplitude oscillation, the following relationships exists.

$$(\left|\overline{R_d}\right| - rd - rc)\, r_c C_c \approx .$$

$$(rd\, r_c C_c / L_d - r_c C_c / \left|\overline{R_d}\right| \overline{C_d} + 1)(r_c \left|\overline{R_d}\right| \overline{C_d} + \left|\overline{R_d}\right| r_c C_c +$$

$$r_d \left|\overline{R_d}\right| \overline{C_d} - L_d - rdr_c C_c) \qquad (23)$$

and

$$\omega_r \approx \left[\frac{\left|\overline{R_d}\right| - r_d - r_c}{L_d \left|\overline{R_d}\right| \overline{C_d} + rc\, C_c (\left|\overline{R_d}\right| \overline{C_d}\, r_d - L_d)} \right]^{1/2} \qquad (24)$$

A proper choice of parameters can lead to an oscillation frequency higher than the maximum self-resonance frequency of the diode. Of course, the cut-off frequency of the diode must be higher than the self-resonance frequency.

The authors then deal with oscillators using distributed circuits. In the article, Figures 12 and 13 show tunnel diode oscillators using straight and re-entrant strip transmission line resonators, respectively. A summary of the performance of several oscillator circuits is given in Table #2 using germanium and gallium-arsenide tunnel diodes.

TABLE II - CHARACTERISTICS OF EXPERIMENTAL TUNNEL DIODE OSCILLATORS

Type of Circuit	Diode Peak Current (ma)	Diode Material	Number of Diodes	Frequency of Oscillation (Mc)	Power Output (mw)	Tuning Range (Mc)	1/8 (Ip-Iv) (Vv-Vp) (mw)
Straight Strip Cavity	210	GaAs	1	610	10.0		8.6
Straight Strip Cavity	53	GaAs	1	900	1.7		2.4
Straight Strip Cavity	50	GaAs	2	950	3.0		4.8
Re-entrant - Strip Cavity	210	GaAs	1	1600	2.0		8.6
Ridge Waveguide Cavity	37	Ge	1	2800	0.7	2700-2900	1.3
Ridge Waveguide Cavity	37	Ge	1	5500	0.2	5400-5600	1.3
Strip Line Re-entrant Cavity	13.7	Ge	1	7130	0.012		0.5
Strip Waveguide	13.7	Ge	1	8350			0.5

Before their conclusion, the authors then state that: "There is little doubt that power outputs an order of magnitude higher than those listed in Table II can be obtained in the near future. Methods that can be used to increase the power output include: paralleling a large number of individual oscillators, use of a number of diodes in a single oscillator circuit, and the use of high current tunnel diodes with distributed junctions. Also research on semiconductor materials may permit the development of high voltage tunnel diodes. In the conclusion, it is stated that: "Experimental oscillators described in this paper have produced power outputs of 10 mw at 610 Mc, 2.0 mw at 1600 Mc, 0.67 mw at 2800 Mc, 0.2 mw at 5500 Mc and 0.01 mw at 7100 Mc. It is anticipated that considerably higher power outputs will be obtained in the near future. Tunnel diode oscillators are compact and rugged, are relatively insensitive to nuclear radiation, have very modest power supply requirements, and can be easily tuned by either mechanical or electrical means; thus, they have significant advantages over lower power vacuum tube oscillators.

Comments on Mr. Sterzer's, et al Paper:

In this paper the authors start from the basic parameters through a small signal analysis to the steady-state oscillation expressions. Several important points were raised; first, C_d and R_d are bias sensitive and r_d is subject to skin effect and hence its real contribution must be evaluated at the frequency of interest. This makes both the cut-off and self-resonant frequencies bias dependant. Secondly, when oscillations expand from their initial response to their steady-state, the frequency changes since C_d and especially R_d is non-linear and hence in the steady-state oscillation condition an average value of C_d and $|R_d|$, $\overline{C_d}$ and $|\overline{R_d}|$ respectively, must be considered. Subsequently, it is stated that to operate above self-resonance (providing there is real negative resistance left or as has been stated, that one operates below the cutoff frequency) since the device looks inductive, the external circuit must be capacitive. The significance of this work is that one can thus take high peak current (Ip) units having large g_d (small R_d) and hence high capacitance and obtain appreciable oscillator power at microwave frequencies.

(9) "HIGH FREQUENCY AND HIGH POWER OPERATION OF TUNNEL DIODES"

by

C. S. Kim and J. B. Hopkins
Digest of Technical Papers, 1961 Int. Solid State Conf., pp. 22 & 23

Representing a tunnel diode by the well known equivalent circuit, the cut off frequency (f_c) and the self-resonance frequency f_r are given by: [1]

$$f_c = \frac{1}{2\pi} \quad \frac{g}{c} \sqrt{\frac{1}{R_s\, g} - 1} \qquad \text{and}$$

$$f_r = \frac{1}{2\pi} \sqrt{\frac{1}{LC} - \frac{g^2}{C^2}}$$

The voltage characteristics of a tunnel diode is determined by the energy level of the junction material and hence is essentially fixed for a particular material. To increase the power output it is necessary to increase the diode current, i. e. the junction area. Assuming g & c are proportional and R_s is inversely proportional to the junction area, f_c is independant of the junction area. On the other hand, f_r^2 is approximately inversely proportional to the junction area, if it is assumed that the inductance L is constant.

To maximize f_r, the smallest possible internal series inductance L_i must be obtained. Therefore, to oscillate at frequencies above self-resonance it is necessary to insert a capacitance [2] C' (or negative inductance L') in series with L_i. The oscillation frequency f_r' ($\omega_r'^2 = \omega_r^2 + 1/Li - L'$; $Li - L' > 0$) can then be extended beyond f_r. The maximum oscillation frequency is ultimately limited by f_c. When the oscillation frequency is made smaller than f_c, the output power increases as shown in Figure #2.

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1 Nelson, D. E, Sterzer, F, "Tunnel Diode Microwave Oscillators with Milliwatt Power Outputs", IRE-WESCON Convention Record, Pt. 1, p. 68, 1960.

2 Kim, C. S., "Circuit Application of Tunnel Diodes", R59ELS111, G-E Internal Publication, p. 35-38, Dec. 1959. Independant work from RCA is published in reference 1.

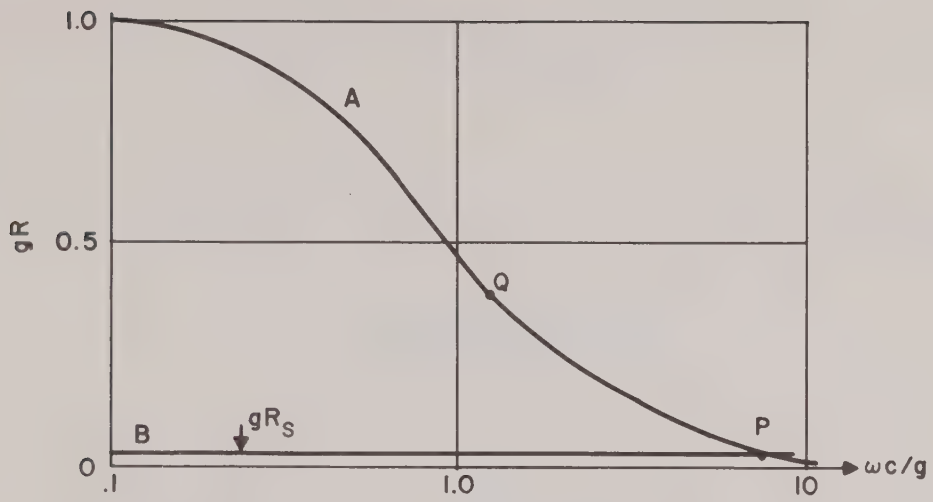


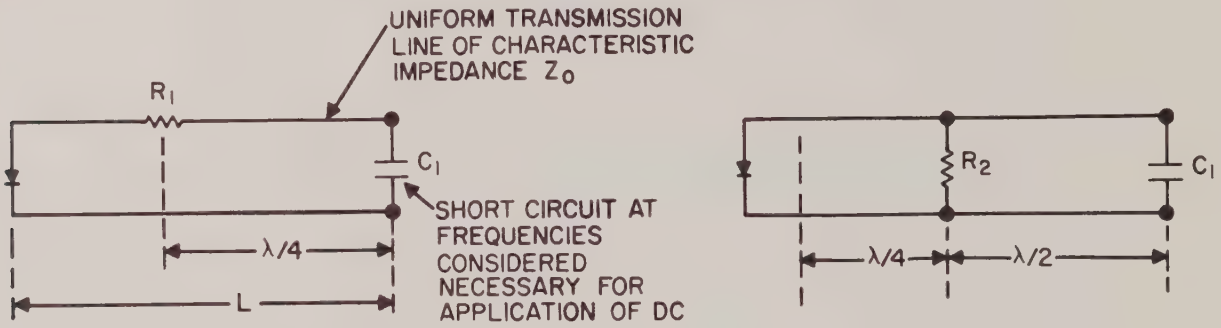
Figure #2 - Plot of g R Versus ωc/g

The oscillation is considered sinusoidal or almost sinusoidal. The power and frequency are normalized to the factor gR and ωc/g, respectively. The power delivered by the negative conductance element divides between R_2 and R_L. The power delivered to R_s is represented by the horizontal line B, and the power delivered to the load $g(R - R_s)$ is the difference between curves A & B. The intersection P, is the cutoff frequency where the output power becomes zero. At microwave frequencies, the capacitance (or negative inductance) element becomes frequency dependant and oscillation will not only take place at f_r', but also at other frequencies.

The negative inductance may be obtained by means of a short-circuited transmission line with length (l) within the range $\lambda/4 < l < \lambda/2$ at the desired oscillation frequency. However at a lower frequency f", the line behaves as an inductance which satisfies another resonance condition, and oscillations can occur simultaneously at two frequencies. This undesirable oscillation at a lower frequency generally becomes the dominant mode, and it can even produce high harmonics. The lower frequency oscillation reduces the magnitude of the average negative conductance of the diode. This reduces the power output at the desired frequency. It is therefore desirable to suppress the undesired oscillation. Figures (3) and (4) show two possible strip-line circuit configurations that will suppress these oscillations. In Figure #3, a resistance R_1 is placed a distance $\lambda/4$ from the short circuited end of the transmission line, where it does not have an appreciable effect at the desired frequency.

However, at other frequencies, especially at lower frequencies, R_1 will be effective in damping out the undesirable oscillations. Figure (4) shows another possible configuration. If R_1 is a dc resistance, R_1 must be smaller than 1/g. Of course, R_1 and R_2 are not restricted to be dc resistances and could be ac impedances with a proper dc value.

Dr. Kim, et al then concludes by describing a cavity type radial transmission line configuration he has used in his work.



Figures #3 and #4 - Circuits to Suppress Lower Frequencies

Comments on Dr. Kim, et al's Paper:

The authors of this article have concentrated on obtaining oscillation above the self-resonance frequency of the diode to obtain higher power outputs at microwave frequencies. They have pointed out the importance and the method to suppress lower frequency-undesired oscillations. The plot in Figure #2 permits an appreciation of the power delivered to the load versus frequency.

(10) "ESAKI DIODE OSCILLATORS FROM 3 TO 40 K MC"

by

R. Trambarulo and C. A. Burrus
IRE Proceedings, October 1960, pp. 1776-1777

"Esaki Diode oscillators operating at microwave frequencies of 10 K mc and below have been reported in the literature. (1-5) Fundamental power up to 40 K mc in frequency has now been obtained from Esaki diode oscillators, with appreciable harmonic output to about 63 K mc. The microwave power at all wavelengths has been sufficient to be detected and displayed by the simplest video methods. Although small area junction diodes of both Gallium Arsenide and Germanium have provided substantial power below 10 K mc, the highest frequencies were obtained with Gallium Arsenide units."

Oscillation ranging in frequency from 2.7 to 33.4 K mc have been obtained in cylindrical cavities. A diode of either GaAs or Ge shunted each cavity at the apex of a re-entrant cone. The cavity resonances occurred at frequencies for which the radius equaled approximately an odd integral number of quarter wavelengths, altered by the diode loading. Power is coupled out of the cavity through a loop. Odd harmonics through the seventh have easily been detected by the authors from a germanium unit having a fundamental frequency of 2.7 K mc. Low order harmonics of much higher fundamental frequency were easily attained with GaAs diodes.

The authors later comment that fundamental microwave power was observed at frequencies from 14.5 to 39.9 K mc in various weakly oscillating circuits, using

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(1) H. S. Sommers, Jr., "Tunnel Diodes as High Frequency Devices". Proc. IRE, Vol. 47, pp. 1201-1206, July 1959.

(2) R. F. Rutz, "A 3000 Mc Lumped Parameter Oscillator using an Esaki Negative Resistance Diode, IBM J. Res. & Dev., Vol. 3, pp. 372-374, Oct. 1959.

(3) R. N. Hall, "Tunnel Diodes", IRE Transactions on Electron Devices, Vol. ED-7, pp. 1-9, Jan. 1960.

(4) R. L. Batdorf, G. C. Dacey, R. L. Wallace and D. J. Walsh, "Esaki Diode in In Sb", J. Appl. Physics, Vol. 31, pp. 613-614, March 1960.

diodes with Ip in the 300 - 700 microamps range. The maximum power measured was 3.5 μ watt in the 17 - 25 K mc region, falling off to about 0.2 μw at 37 K mc. The maximum fundamental power observed with strongly oscillating circuits was 50 μw at 16.7 K mc. Appreciable harmonic output was observed to 62.7 K mc, where the SNR of the detected output displayed on our oscilloscope was as high as 20:1.

Because both the negative resistance and capacitance of a diode varies with voltage, the frequency of oscillation depends somewhat upon bias, and the voltage tuning range may be as much as several per cent of the center frequency. As the bias voltage is increased, the frequency of weakly oscillating circuits decreases monotonically. For large amplitudes of oscillation, however, the frequency may go through a minimum or maximum or both.

The authors then state that they found the performance of GaAs diodes to be superior at high frequencies to that of diodes made from the most highly doped germanium available to them. GaAs diodes loaded a cavity resonant at 4 K mc almost inperceptibly, but the best Ge units in the same circuit oscillated barely in excess of 2.8 K mc. The higher fundamental frequencies and harmonics have been obtained with GaAs units. It appears that these GaAs units have a lower barrier capacitance than Ge units of the same negative resistance, and show promise for application in amplifiers and oscillators well into the millimeter wavelength region.

Comments on the above paper:

Mr. Trambarulo et al. have shown the possibility of operating micropower oscillators in the fundamental mode up to 39.9 K mc and in the harmonic mode up to about 63 K mc. It was further shown that the frequency is bias sensitive and a tuning range of several percent of the center frequency can be achieved. The authors then express their opinion on the relative merits of GaAs and Ge units in the higher frequency (millimeter) cases. They observed GaAs to have lower barrier capacitance for a given negative resistance.

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(5) J. K. Pulfer, "Voltage Tuning in Tunnel Diode Oscillators", Proc. IRE, Vol. 48, p. 1155, June 1960.

(11) "MILLIMETER WAVE ESAKI DIODE OSCILLATORS"

by

C. A. Burrus
IRE Proceedings, December 1960, p. 2024

Mr. Burrus makes mention of recent reports of oscillators operating up to 40 K mc (7.5 mm wavelength) (1) Then he says that more recently the high frequency limit at which fundamental oscillation have been observed has been extended to 103 K mc (2.9 mm). The Esaki diodes used in this latter experiment were made by electrically "forming" a point contact between zinc and heavily doped (ρ = 0.0007 - 0.0009 ohm-cm) n-type GaAs. Peak current of these units ranged from a few hundred microamps to several milliamps with Ip/Iv of > 3:1.

The highest frequency fundamental oscillation observed to date was at 103 K mc. The diode had a peak current of 600 μa and was mounted in a 0.122 x 0.010 inch waveguide (48.4 K mc low frequency cutoff). Oscillation occurred from 93 to 103 K mc, tunable over this range by movement of the waveguide piston.

In a few instances the frequency could be varied 33 to 38 per cent by movement of the waveguide piston. The frequency variation more commonly observed by such tuning, however, was 10 to 25 per cent. The frequency sweep provided by the diode bias variation was usually 0.5 to 2.0 per cent.

Diodes employed in the 50 to 100 K mc oscillators had peak currents in the 0.5 to 2.0 ma range. Diodes having lower peak currents did not oscillate in the circuits employed, and those with higher peak currents usually oscillated at frequencies below the waveguide cutoff so that only harmonic output was observed.

Some power measurements were made using a video detector diode calibrated at 58 K mc against a colorimeter. (4) On this basis, the maximum power so far obtained was 25 μw, or approximately - 16 dbm at 50 K mc, falling off to about 2 μw at 90 K mc and a few tenths of a microwatt at 100 K mc.

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(1) R. F. Trambarulo and C. A. Burrus, "Esaki Diode Oscillators from 3 to 40 K mc", Proc. IRE Vol. 48, pp. 1776-1777, October 1960.

(4) W. M. Shapeless, "A Colorimeter for Power Measurements at Millimeter Wavelengths" IRE Transactions or Microwave Theory & Techniques, Vol. MTT-2, pp. 45-47, September 1954.

The author expects that the high frequency limit of oscillation, the operating efficiency and the tuning range of the oscillators might be increased with more nearly optimum circuitry.

Comments on the paper by Mr. Burrus:

In the above paper, Mr. Burrus has shown oscillations up to 103 K mc (2.9 mm). The diodes are "electrically formed point contact" units using zinc and n-type GaAs. These units have been made with peak currents of a few hundred microamps to several milliamps. The oscillator frequency achieved in a waveguide could be tuned 10 - 25% of the center frequency by a waveguide piston (33 - 38% in a few instances) and 0.5 to 2.0% by changing the bias voltage. At 50 K mc - 25 μw (-16 dbm) of power was measured, falling off to 2 μw at 90 K mc.

(12) "VOLTAGE TUNING IN TUNNEL DIODE OSCILLATORS"

by

J. K. Pulfer
IRE Proceedings, June 1960, p. 1155

During some recent experimental work on microwave tunnel diode oscillators, voltage tuning ranges of up to 12 per cent have been obtained by varying the operating bias of the oscillator. Power output was approximately - 30 dbm.

Figure #1 shows a curve of frequency vs. voltage obtained with an oscillator using an Esaki diode, manufactured by the Sony Corporation of Japan. The diode was mounted at one end of a low impedance coaxial line, as shown in Figure #2, and the other end of the line was terminated in a matched load, across which the operating voltage was applied.

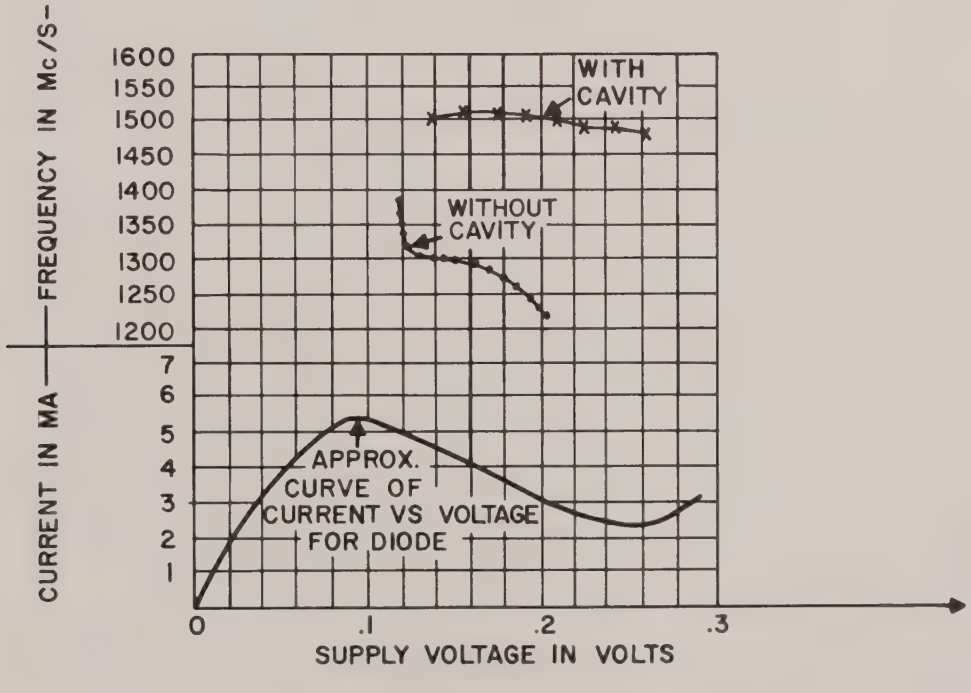

Figure #1 - Voltage Tuning Curves for T. D. Oscillator

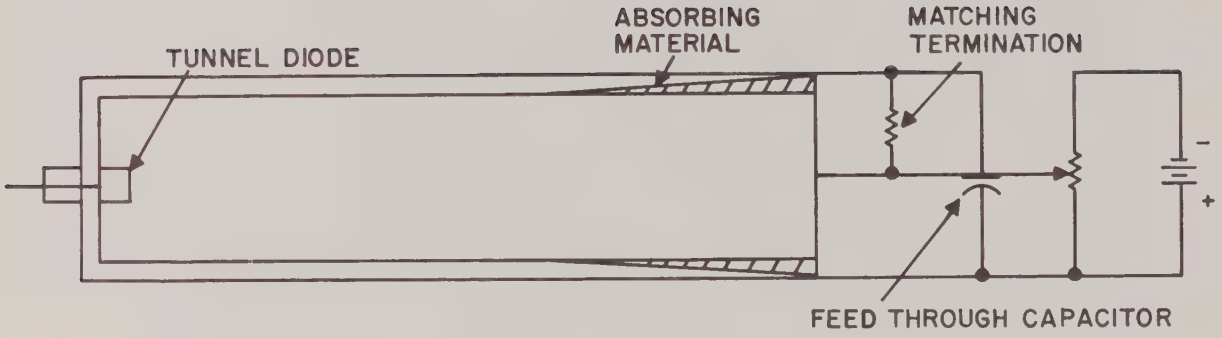

Figure #2 - Tunnel Diode Oscillator (output was taken from a probe at the diode end).

By varying the operating voltage, the effective negative resistance of the diode at which the oscillation reach a steady state may be varied, thus changing the frequency. The relative amplitude and phase of the harmonics generated will also vary with supply voltage, and, this also affects the oscillation frequency. The nonlinearity of frequency vs. voltage is thought to be due to resonances at the harmonic frequencies.

Comments on Mr. Pulfer's Paper:

Mr. Pulfer has shown a circuit and data on the effect of bias tuning a microwave oscillator. A tuning range of 12 per cent was achieved.

SECTION III

CONVERTERS

ABSTRACT:

The Papers in this section deal with two distinct classes of converters. Sterzer et al. discusses the autodyne (self-oscillating) converter while the others deal with the driven mixer type. Kim's analytical paper covers both classes of operation - further broken down into the fundamental and harmonic case - in the general form.

Basically much of the information derived in the amplifier and oscillator sections will hold true for this section. The Specifications of the tunnel diode to be used do not change radically.

In the driven mixer case the noise figure is said to be greater than in the amplifier configuration. As the gain of the mixer is increased this seems to become less true. Kim states that when operating well below "cut-off", less than 3 db noise figure can theoretically be obtained. Chang, Breitzer and Robertson locate their operating point in the positive conductance region, somewhat below Ip, while Dickens biases at the minimum Io/gd point (and seems to achieve lower noise figures). Sterzer naturally operates in the negative conductance region since the circuit must "self-oscillate". Sterzer draws some conclusions on oscillator pulling by saying that the signal level at which conversion gain decreases is proportional to Ip. At higher signal levels oscillator pulling occurs. Robertson emphasizes coupling stability by using two matched diodes in a Hybrid Coupler arrangement.

Performance data on the converter circuits discussed in this section is shown on the following page.

Author	Sig. Freq.	Osc. Freq.	I. F.	O. P.	CG	BW	N. F.	Diode Used
K. K. N. Chang etal	210 Mc	240 Mc	30 Mc	18 ma 29 mv	6 db	0. 9 Mc	5. 2 db	Ge 35 ma
K. K. N. Chang etal	210 Mc	240 Mc	30 Mc	20 ma 60 mv	22. 7 db	0. 15 Mc	2. 8 db	GaAs 23. 6 m
Breitzer	-	-	-	100 mv	-	-	8. 3 db	Ge 1N2940
Dickens et al	1200 Mc	1170 Mc	30 Mc	min Io/gd)	26 db	4. 0 Mc	4. 7 db	Ge 1N3218A
Robertson	400 Mc	-	30 Mc	Pos. Cond. region near Ip	10 db	-	7. 0 db#	Ge 1N2939
Sterzer et al	1665 Mc	-	30 Mc			+2 dbm	-	Ge Ip = 5 ma

two channel noise figure

(13) "LOW NOISE TUNNEL DIODE DOWN CONVERTER HAVING CONVERSION GAIN"

by

K. K. N. Chang
IRE Proceedings, May 1960, pp. 854-858

In his introduction, Mr. Chang makes the following statement: "The low noise, high-gain, conversion of a high-frequency signal to a lower frequency signal has not been possible with previous mixer-devices. Ordinary crystal mixers, which make use of the nonlinearity of their positive resistance, exhibit a conversion loss and a poor noise factor. The recently introduced parametric converters, which operate on a nonlinear capacitance or inductance basis, have achieved good noise factors with up-conversion gain. The parametric down-converters, however, have poor noise factors. For such down-converters, it is found that the excess noise factor (i. e., the noise factor minus unity) varies roughly as the ratio of the input frequency to the output frequency. Thus, for a ten to one frequency down conversion, it is almost impossible to convert a microwave frequency into a low intermediate frequency with a reasonable noise factor by parametric converters.

The purpose of this paper is to report on a down-converter using a tunnel diode (Esaki diode)[1, 2] as the nonlinear resistance element. The fact that the negative resistance characteristic of a tunnel diode can be utilized to achieve low-noise amplification has already been demonstrated. (3) It is shown that the nonlinearity of this negative resistance characteristic can be used for frequency conversion. Since the nonlinearity of a resistance, not a reactance, is utilized for frequency conversion, the noise factor is independent of the ratio of the input frequency to the output frequency. Thus, by using a tunnel diode, a low-noise down-converter with conversion gain has been achieved."

The circuit that Mr. Chang shows in his paper is illustrated in Figure #1.

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(1) L. Esaki, "New Phenonenon is narrow ge pn junctions", Phys. Rev. Vol. 109, p. 603 (L); 1958.

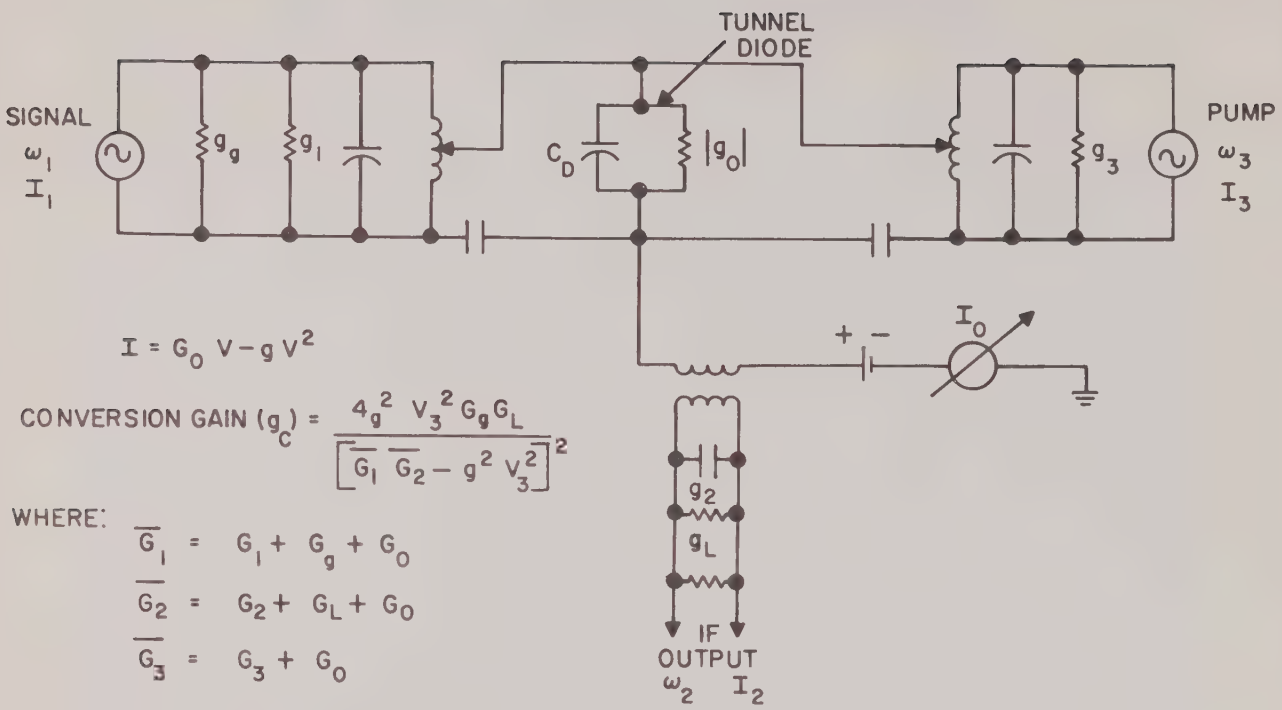
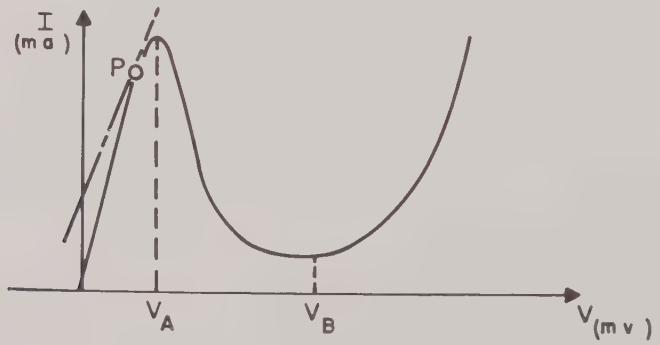


Figure #1 - Schematic Diagram of Converter Circuit

The D. C. operating point chosen by Mr. Chang is shown in Figure #2



NOTE: for conversion gain oscillator voltage V_3 must be sufficiently large to swing the current below the value of the current at the operating point (in the negative) conductance region.

Figure #2 - Tunnel Diode I-V Characteristic and Location of Operating Point "P" chosen by Mr. Chang.

(2) H. S. Sommers, Jr. "Tunnel Diodes as High Frequency Devices", Proc. IRE, Vol. 47, pp. 1201-1206, July 1959.

(3) K. K. N. Chang, "Low Noise Tunnel Diode Amplifier", Proc. IRE, Vol. 47, pp. 1268-1269, July 1959.

The noise figure equation derived by Mr. Chang is:

$$F = 1 + \frac{T}{To}\left[\frac{Ge}{Gg} + \frac{Gl}{Gg} + \frac{(G_L + G_2 + G_e)\,\overline{Gl}}{Gg\;\overline{G_2}}\;\;\frac{1}{1 - \frac{2\left(\sqrt{1 + \overline{Gc}} - 1\right)}{\overline{Gc}}}\right]$$

The tests were conducted by Mr. Chang with a 35 ma Ip (Vp = 62 mv) germanium tunnel diode operated at 18 ma and 29 mv yielding a positive Go = 0.42 mho, and a 23.6 ma Ip (Vp = 100 mv) GaAs unit operated at 20 ma and 60 mv with a positive Go = 0.20 mho.

The operating frequencies: f_1 = 210 mc/s, f_2 = 30 mc/s and f_3 = 240 mc/s.

The results are shown in Table I: (See next page)

TABLE I

DIODE	P. G.		B. W.		N. F.		SENSITIVITY
	Measured	Computed	Measured	Computed	Measured	Computed	Measured
Germanium:	6.0 db	6.0 db	0.9 Mc	0.6 Mc	5.2 db	4.4 db	1.5 μv
GaAs:	22.7 db	22.7 db	0.15 Mc	0.26 Mc	2.8 db	3.8 db	0.25 μv

Mr. Chang concludes:

"The present experiments have demonstrated that the tunnel diode can be utilized as a low-noise, high-gain down converter in the UHF range. It seems feasible to extend the operating frequencies into the microwave region. The diodes for such applications should be designed with as low a capacitance as possible. The associated circuits, and especially the input circuit, should have high conductance and low Q's in order to realize optimum bandwidth and noise figures".

Comments on Mr. Chang's article:

Mr. Chang has investigated a driven mixer circuit having an operating point in the positive conductance region somewhat below the peak point. The separate local-oscillator must furnish enough oscillator power to the mixer to swing into the negative conductance region to a value of current lower than the D. C. bias current.

The results obtained by Mr. Chang show gallium arsenide units to be superior to germanium in terms of gain and noise figure.

Mr. Chang feels that self-oscillating mixers (autodyne converters) are possible and that with appropriate diodes the operating frequencies can be extended into the microwave region.

(14) "NOISE FIGURE OF TUNNEL DIODE MIXER"

by

D. I. Breitzer
IRE Proceedings, May 1960, pp. 935-936

Mr. Breitzer has made an analysis of a driven tunnel diode mixer. His stated assumptions were:

1. The diode sees a short circuit at the image frequency.
2. The only source of noise in the diode is the shot noise associated with the tunnel current.
3. The series resistance of the diode and the losses of the input and output tuned circuits are negligible, and
4. The diode capacitance is constant.

Because of the second assumption, his consideration was limited to that portion of the I-V characteristic of the diode extending from reverse voltages up to, or somewhat beyond, the point of maximum negative conductance. The valley region is especially excluded because of the excess noise.

Mr. Breitzer states that at midband the conversion gain (of a nonlinear resistance type diode) defined as the ratio of I. F. output power to the power available from the RF source, is given as:

$$G = \frac{4 r_1 r_2 g_1^2}{\left[(1 + g_o r_1)(1 + g_o r_2) - g_1^2 r_1 r_2\right]^2} \qquad (1)$$

where, r_1 and r_2 are the resistances seen by the diode at R. F. and I. F., respectively, g_o is the average conductance of the diode over the local oscillator cycle and g_1 is the conversion conductance of the diode. The values of g_o and g_1 are obtained from the Fourrier series representation of the instantaneous incremental conductance g of the diode considered as a function of the local oscillator phase θ. Thus:

$$g(\theta) = g_o + g_1 \cos \theta + 2 g_2 \cos 2\theta + \ldots\ldots\ldots \qquad (2)$$

If $g(\theta)$ is restricted to non-negative values than $0 \leq |g_1| \leq g_o$ and the maximum possible value of G (gain) for positive r_1 and r_2 is unity.

However, if g (θ) is free to assume negative values, then $g_o < |g_1|$ is possible and the conversion gain cannot only exceed unity, but will become infinite for

$$r_2 = r_{2c} = \frac{1 + g_o r_1}{g_1^2 r_1 - g_o (1 + g_o r_1)} \qquad (3)$$

Mr. Breitzer continues:

It can be shown that for stable operation the quantities $(1 + g_o r_1)$ and $(1 + g_o r_2)$ must both be positive or both negative. Assuming that both are positive, the condition that r_{2c} be positive is that

$$\gamma = \frac{|g_1| r_1}{1 + g_o r_1} > \frac{g_o}{|g_1|} = \alpha \qquad (4)$$

For values of r_2 slightly below r_{2c}, the mixer gain is similar to that of a negative resistance amplifier, i. e., the product of voltage gain and bandwidth is constant. The noise figure of the tunnel diode mixer is:

(5)

$$F = 1 + \frac{q(I_{of} + I_{or})}{2k T_1 |g_1|} \frac{1 + \gamma^2 \pm 2K\gamma}{\gamma(1 - \alpha\gamma)} + \frac{1}{\gamma(1 - \alpha\gamma)|g_1| r_2} \frac{T2}{T1}$$

According to equation #5, the larger r_2 the better the noise figure of the mixer. However, the mixer conversion gain is more critically dependant on r_2 and high conversion gain is desirable to minimize the over-all noise figure of the system.

Mr. Breitzer than gives some results calculated for a General Electric ZJ56 (1N2940). The calculation was made for various values of bias and local oscillator drive on the assumption that the local oscillator voltage across the tunnel diode is sinusoidal. The best value obtained for the noise figure was 8. 3 db for a forward bias of 100 mv and a peak local oscillator voltage of 90 mv. For this set of operating conditions g_o = + 0. 0029 mho, g_1 + - 0. 0069 mho, I_o = 0. 924 ma, I_1 = 0. 346 ma, r_1 opt. = 135 ohms and r_{2c} = 600 ohms.

By way of comparison, the same diode used as a negative-resistance amplifier has a noise figure of about 5 db. The noise figure as a mixer is thus

poorer than the noise figure as an amplifier. This result is to be expected unless more effective use can be made of "noise compensation" in the mixer. [4] Unfortunately, to obtain significant noise compensation, Class C operation of the diode is required and this mode of operation is not possible for the tunnel diode.

Comments on Mr. Breitzer's article:

Mr. Breitzer seems to come to somewhat different conclusions than Mr. Chang (in the previous article) as to the optimum choice of the operating point.

A noteworthy conclusion: the tunnel diode as a driven mixer is noisier than when used as a negative resistance amplifier.

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(4) A Van der Ziel "Noise" Prentice-Hall Inc., New York, New York, pp. 249-250; 1954.

(15) "TUNNEL DIODE CONVERTER ANALYSIS"

by

C. S. Kim
Electronic Laboratory
General Electric Company
Syracuse, New York

C. S. Kim analyses tunnel diode converters for two different conditions of operation. First, in the self-oscillating (autodyne) case and secondly, in the driven mixer case. He then subdivides these two into two further categories:

1) the fundamental converter

2) the harmonic converter.

His general analysis also includes the effect of series resistance (R_s), series inductance (L_s) and shunt capacitance (C_D) as well as the effect of the image termination. He discusses the choise of several bias operating points. Although this paper is analytical in nature and does not include measured data, it is a thorough study of the problem and the author does conclude by saying that by operating at a frequency much lower than the cutoff frequency (f_{ro}) a 3 db noise figure seems obtainable for the self oscillating converter type.

NOTE: Due to the analytical aspects of Dr. Kim's article, it would be difficult to abstract it. However, because of it's importance, we have reproduced the article. It is available to you upon request by writing to: Advertising & Sales Promotion, Semiconductor Products Department, General Electric Company, Kelley Building, Liverpool, New York.

THE JOHNS HOPKINS UNIVERSITY
RADIATION LABORATORY
BALTIMORE, MARYLAND

(16) "A TUNNEL DIODE AMPLIFYING CONVERTER"

by

L. E. Dickens and C. R. Gneiting

NOTE: The enclosed material, while being copied excerpts from another report, is complete as to the material published in the January issue of the Transactions on Microwave Theory and Techniques.

This research was supported by the United States Air Force through WWRNGW of the Wright Air Development Division of the Air Research and Development Command.

REPORT: That tunnel diodes have a bright future as relatively low noise, low power consumption, amplifying devices has already been indicated in the published literature [13, 14, 15, 16]. During the past quarter the use of the tunnel diode as a mixing element has been the subject of a study, certain results of which are given below.

In the usual microwave or VHF receiver superheterodyne principles are most often utilized. In such a system the signal is received by some form of antenna, amplified, and then transmitted to a mixer, or transmitted directly to the mixer without amplification. In the mixer the received signal and the local oscillator signal operate usually on a nonlinear variable resistance device from which the intermediate frequency signal is then derived.

This mixer is usually quite noisy and quite lossy; the diode noise figures being on the order of 7.0 db and the diode losses being on the order of 6.0 db.

It is possible to place a negative resistance amplifier before the mixer to minimize the effects of the mixer noise figure and loss; but the peculiar current-voltage (I-V) characteristic of the tunnel diode presents a much more desirable solution [17]. The RF amplifier and variable resistance mixer may be entirely eliminated, and in their place the tunnel diode may be used to obtain the necessary high gain and low noise figure and to convert from the signal frequency down to the intermediate frequency.

---

(13) Sommers, H. S., Jr., "Tunnel Diodes as High-Frequency Devices," Proc. IRE, Vol. 47, No. 7, pp. 1201-1206, July 1959.

(14) Chang, K. K. N., "Low Noise Tunnel Diode Amplifier," Proc. IRE, Vol. 47, No. 7, p. 1268, July 1959.

(15) Lesk, I. A., Halonyak, N., Jr., Davidsohn, U. S., and Aarons, M. W., "Germanium and Silicon Tunnel Diodes -- Design, Operation and Application," 1959 IRE Wescon Convention Record, Part 3, Electron Devices, pp. 9-31.

(16) Hines, M. E., "High Frequency Negative Resistance Circuit Principles for Esaki Diode Applications," Bell System Tech. J., Vol. 34, No. 3, pp. 477-513, May 1960.

(17) Chang, K. K. N., Heilmeier, G. H., and Prager, H. J., "Low Noise Tunnel Diode Down Converter Having Conversion Gain," Proc. IRE, Vol. 48, No. 5, pp. 854-858, May 1960.

Figure 1 is representative of the I-V characteristic of the diode used in the experiments. Figure 2 is an oscilloscope presentation of the conductance of the same diode versus voltage.

The tunnel diode is used as the mixing diode; the bias is adjusted for stable operation at point A. The total dc source resistance (r_1) must be less than the magnitude of the negative resistance of the diode (point B).

The local oscillator (LO) signal is coupled into the mixer by the usual techniques. Sufficient LO drive must be supplied to modulate the tunnel diode about the I_o - V_o point. The LO is decoupled sufficiently so that it loads neither the RF nor IF circuit.

The RF equivalent circuit used to define the device parameters is shown in Figure 3(a). Figure 3(b) shows the equivalent circuit of the complete mixer as used in the experiments. $\vec{G}$ is the exponential Fourier series representation of the conductance waveform under local oscillator modulation. $\vec{I}$ is the exponential Fourier series representation of the diode current under LO modulation. F_2 is a band-pass filter arranged to pass the input signals from the generator, and to act as an open circuit for the image signals. F_1 is a low-pass filter which passes the IF signals but blocks the RF input signals.

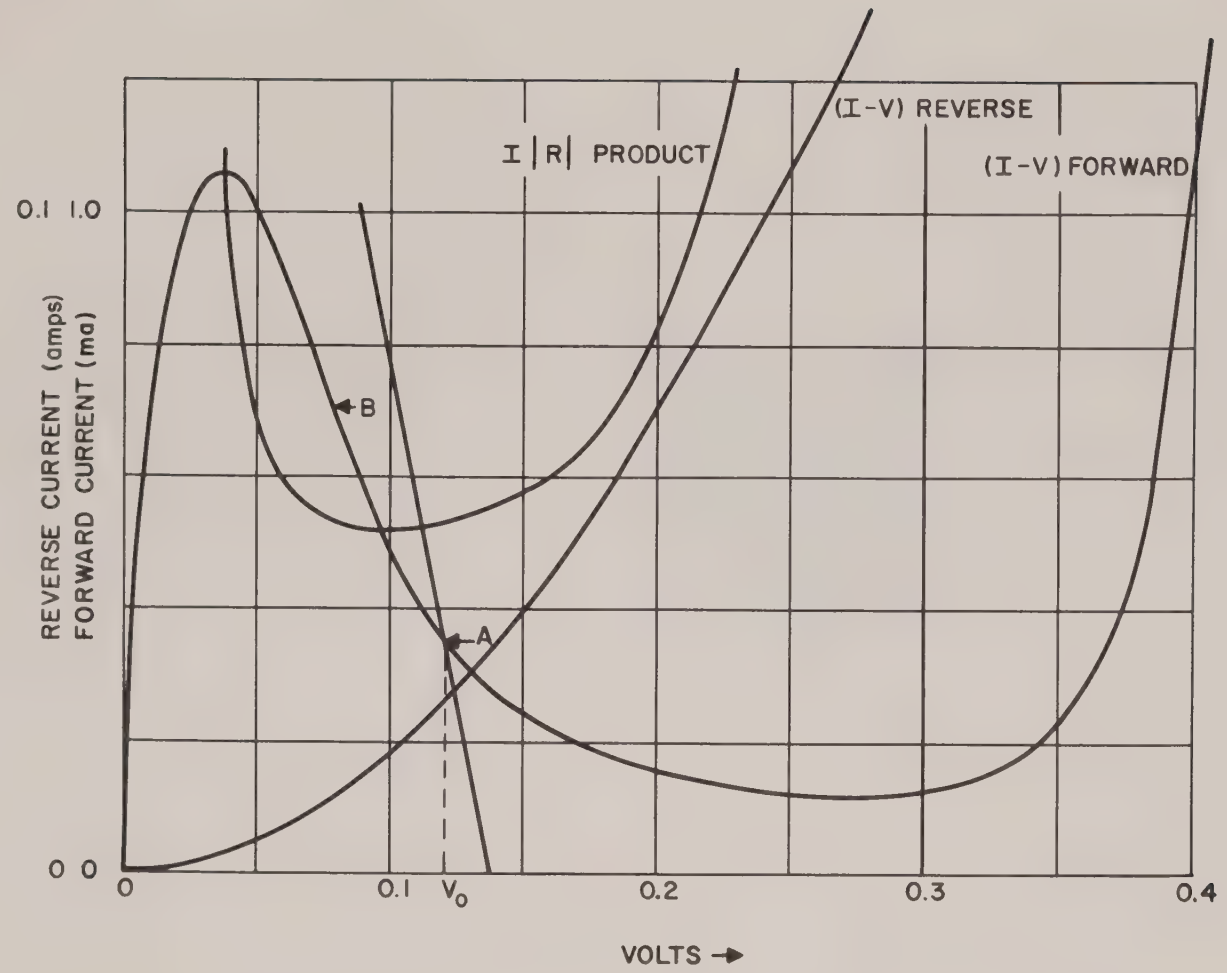


FIGURE #1 - CURRENT-VOLTAGE CHARACTERISTIC OF THE GERMANIUM DIODE USED IN THE CALCULATIONS

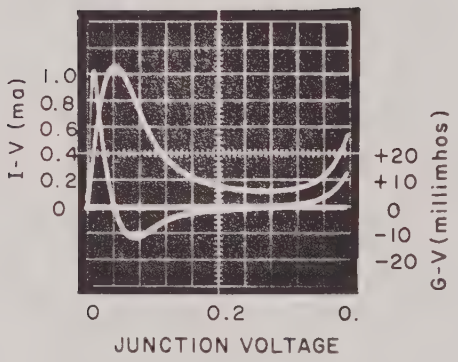
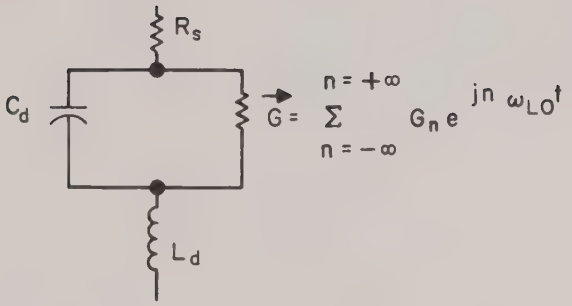
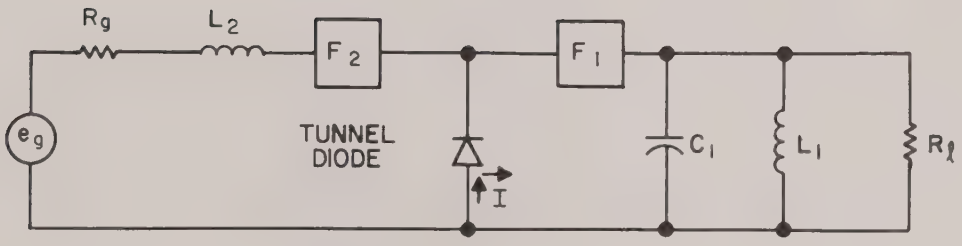


FIGURE #2 - OSCILLOSCOPE PRESENTATION OF THE CONDUCTANCE VS VOLTAGE CHARACTERISTICS OF THE DIODE USED IN THE EXPERIMENTS.



(a) RF EQUIVALENT CIRCUIT OF TUNNEL DIODE



(b) CIRCUIT OF THE TUNNEL DIODE MIXER

FIGURE 3

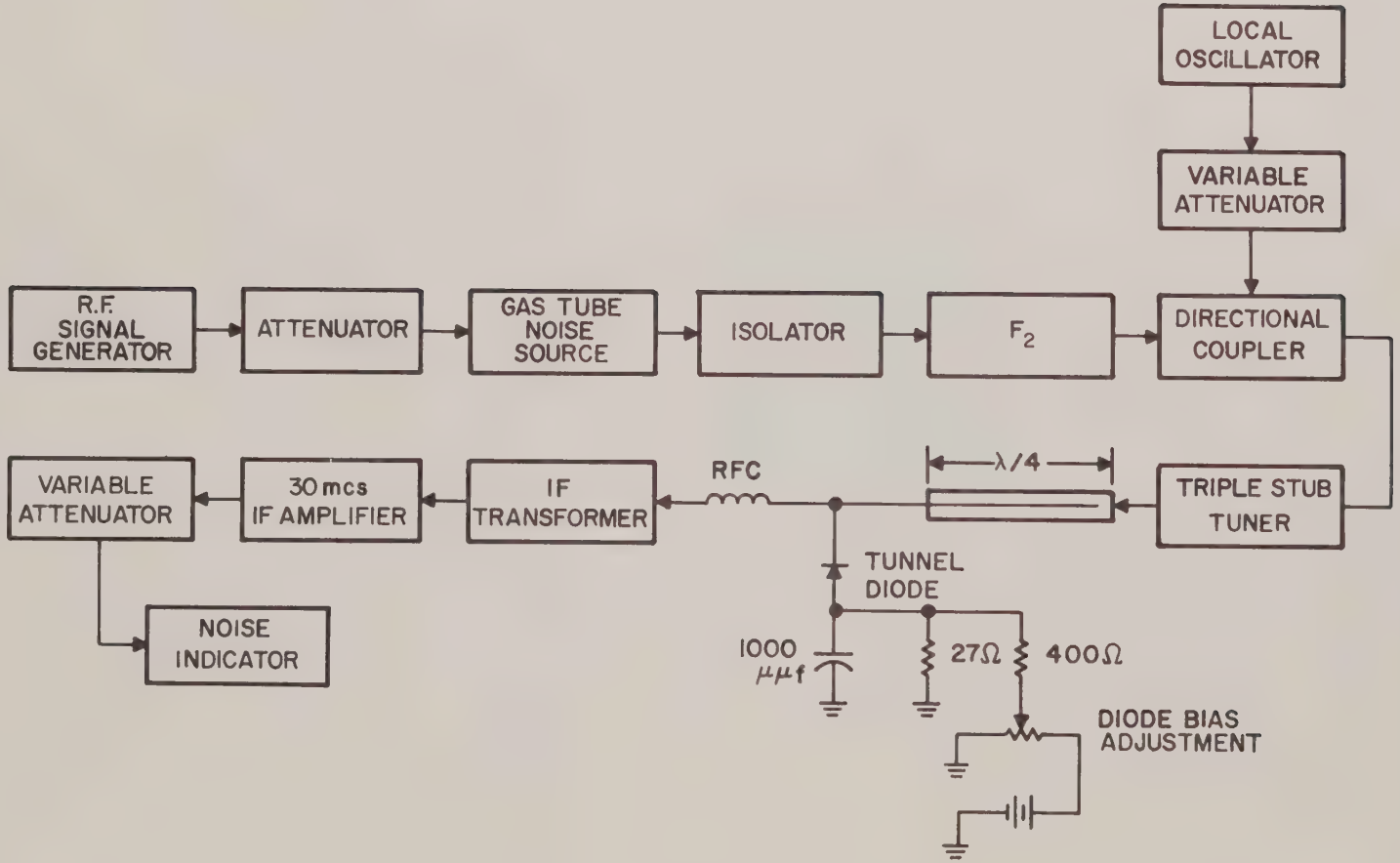


FIGURE 4 - BLOCK DIAGRAM OF THE EXPERIMENTAL SYSTEM

With the circuit configuration as shown in Figure 3, let the following quantities be defined.

$$Q_2 = \frac{\omega_2 (L_2 + L_d)}{R_g + R_s} \; ; \; \overline{Q_1} = \frac{\omega_1 \; C_1}{(G_1 + G_o)} \; ; \; \overline{Q_2} = \frac{\omega_2 \; C_d}{(G_g + G_s + G_o)} \tag{1}$$

$$G_g + G_s = \frac{1}{(R_g + R_s)(1 + Q_2^2)} \tag{2}$$

$$\frac{G_s}{G_g} = \frac{R_s}{R_g} \tag{3}$$

where

(1) R_g is the series source resistance and G_g is the equivalent source conductance

(2) R_s is the series loss resistance and G_s is the equivalent shunt loss conductance

(3) $R_s << R_g$; $L_d << L_1$

(4) $\omega_1 = 2\pi$ (intermediate frequency)

$\omega_2 = 2\pi$ (radio frequency).

Based on the above definitions for the equivalent conductances, the conversion power gain (W_p) may be shown to be as in Equation (4).

$$W_p = \frac{G_1 \; G_g}{G_1^2} \cdot \frac{4a^2}{(1 - a)^2} \; , \tag{4}$$

where

$$a = \frac{G_1^2}{(G_1 + G_o)(G_g + G_s + G_o)} \tag{5}$$

and the bandwidth (B) may be found from the following relation, Equation (6)

$$B = \frac{\omega_1 (1 - a)}{\overline{Q_1} + \overline{Q_2} \left(\frac{\omega_1}{\omega_2} \right)} \tag{6}$$

The noise figure is given by Equation (7).

$$F = 1 + \frac{T}{T_o}\left[\frac{G_s}{G_g} + \frac{G_{eq0}}{G_g} - \frac{G_{eq1}}{G_1}\left(\frac{G_g + G_s + G_o}{G_g}\right) + \frac{\left(G_{eq0} + \frac{T_1}{T} G_1\right)(G_g + G_s + G_o)^2}{G_g G_1^2}\right] \tag{7}$$

where

T_o = reference (source) temperature in degrees Kelvin

T = temperature of the diode

T_1 = temperature of the load

G_{eq0} = equivalent noise conductance of the diode = $\frac{eI_o}{2KT}$

$G_{eq1} = \frac{e(2I_1)}{2KT}$

where I_o and I_1 are the terms of the Fourier expansion corresponding to n = 0 and n = 1 respectively.

A converter using germanium tunnel diodes has been operated at 1,200 mc, the IF being 30 mc with the local oscillator set at 1,170 mc. The diode had a peak current of 1 ma.

Several noise figure calculations were made for various points of operation about the point of maximum positive slope of the conductance curve in the negative resistance region. Best noise figures were obtained upon biasing so that V_o is between 100 and 200 mv; this corresponds closely to the position of minimum IR product [18] or I/G ratio [19]. With 50 mv of peak LO drive, moderate values of G_1

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(18) Chang, K. K. N., "The Optimum Noise Performance of Tunnel Diode Amplifiers," Proc. IRE, Vol. 48, No. 1, pp. 107-108, January 1960.

(19) Tieman, J. J., "Shot Noise in Tunnel Diode Amplifiers," Proc. IRE, Vol. 48, No. 8, pp. 1418-1423, August 1960.

may be obtained with relatively large negative values of G_o. The best calculated value for the noise figure was about 4.7 db, and was obtained from the following data.

For a bias voltage of 125 mv, and a peak local oscillator voltage of 50 mv, the following parameters are obtained:

$$G_o = -0.007 \text{ mho}; \qquad I_o = 0.408 \text{ ma}; \qquad G_{eq0} = 0.00816 .$$

$$G_1 = -0.0031 \text{ mho}; \qquad I_1 = 0.14 \text{ ma}; \qquad G_{eq1} = 0.0056 .$$

It has been assumed that the LO voltage waveform at the junction remains sinusoidal.

The series resistance, R_s, and the junction capacitance, C_d, are known ($R_s = 1$ ohm; $C_d = 5$ $\mu\mu$f); therefore, the equivalent loss conductance $G_s = 0.0015$ mho. Assuming $\alpha = 0.7$ and the generator conductance $G_g = 0.006$, and using Equation (5), we find G_1 to be $G_1 = 0.0338$ mho. By Equation (4) the gain is found to be about 26 db. Assuming in Equation (7) $\frac{T}{T_o} = \frac{T_1}{T} = 1$, we find the noise figure to be 4.7 db. By Equation (6) the band-pass is calculated to be about 4.0 mc.

A block diagram of the converting system as tested is shown in Figure 4. A ferrite isolator was used at the input to insure good termination for the gas tube noise source and also to isolate the converter from changes in the generator impedance (the differences in the fired and unfired condition of the noise source). F_2 is a cavity band-pass filter which, in conjunction with the triple stub tuner, acted to pass the signal but reflect the image in such a manner as to present an open circuit to the image. The directional coupler serves only the cause of LO injection with good decoupling ($\approx$ 25 db). The quarter wave section shown serves not only to pass the signal, but to block the IF. The IF transformer changes the IF impedance (50 ohms) to the desired value of about 30 ohms.

With the system adjusted as described, it was a relatively easy task to bring the measured noise figure within the range of 6-8 db. By fine adjustments of the tuning on the RF side, the bias point, and the local oscillator drive, the noise figure was brought within the range of the calculated value. With the noise figure brought down to about 4.5 db the measured bandwidth was about 4.0 mc. The signal generator was used to measure the bandwidth and also to insure the absence of image response and other spurious responses that might cause erroneous noise figure indications.

There remains now a question of relative merit of the tunnel diode converter and the tunnel diode amplifier followed by a standard converter. There is no generality that may be applied. The question must be resolved for each separate application, since each application will have a different set of rules governing stability, gain bandwidth, overall noise figure, and so forth. As has been pointed out [20], the noise figure of the converter is in general higher than that of the amplifier. As may be seen from Equations (4) through (7), as the gain is made very large the term $(G_g + G_s + G_o)$ becomes very small, and the noise figure equation reduces to that which applies to the one port negative conductance amplifier. If the high gain converter is used, the system noise figure may approach, or surpass, that of the system using the negative conductance amplifier, since in the process of converting from signal to IF there is no lossy diode mixer involved.

Comments on Messrs. Dickens' and Gneiting's paper:

Messrs. Dickens and Gneiting have analyzed the driven mixer and have stated expressions for Conversion Gain, Bandwidth and noise figure.

Using a 1 ma (Ip) germanium tunnel diode (such as the 1N3218A) they have built and tested a 1200 Mc converter with a 30 Mc IF, (hence the local oscillator was set at 1170 Mc). Operating at the minimum IR or $\frac{I}{G}$ point with a 50 mv L. O. drive they calculated a 4.7 db minimum noise figure and a 4.0 Mc bandwidth for 26 db conversion gain.

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(20) Breitzer, D.I., "Noise Figure of Tunnel Diode Mixer," Proc. IRE, Vol. 48, No. 5, pp. 935-936, May 1960.

(17) "A BROAD-BAND HYBRID COUPLED TUNNEL DIODE DOWN CONVERTER"

by

W. J. Robertson
IRE Proc. Dec. 1960, pp. 2023-2024

"Recent work on tunnel diode traveling-wave amplifiers demonstrated the need for a low-noise broad-band radio frequency mixer that provides isolation between the tunnel diode amplifier and the local oscillator. It was found that the presence of even small amounts of local oscillator signal greatly reduced amplifier gain. A hybrid coupled tunnel diode down converter which has the required isolation, moderate power gain, a very large bandwidth, a reasonable noise figure, and absolute stability has been constructed and tested.

A pair of matched tunnel diodes (G. E. 1N2939) and a balanced I. F. circuit are connected to two arms of a Sage 751 Hybrid as shown in Figure #1. The 30 mc/s transformer consists of a balanced tank circuit with the center tap of the inductor opened so that the tunnel diodes may be individually biased. When biased near the peak current point in the positive conductance region of the I-V characteristic as indicated in Figure #2, the tunnel diodes appear as positive balanced loads to the local oscillator and hybrid action causes cancellation of the local oscillator signal at the signal input terminals.

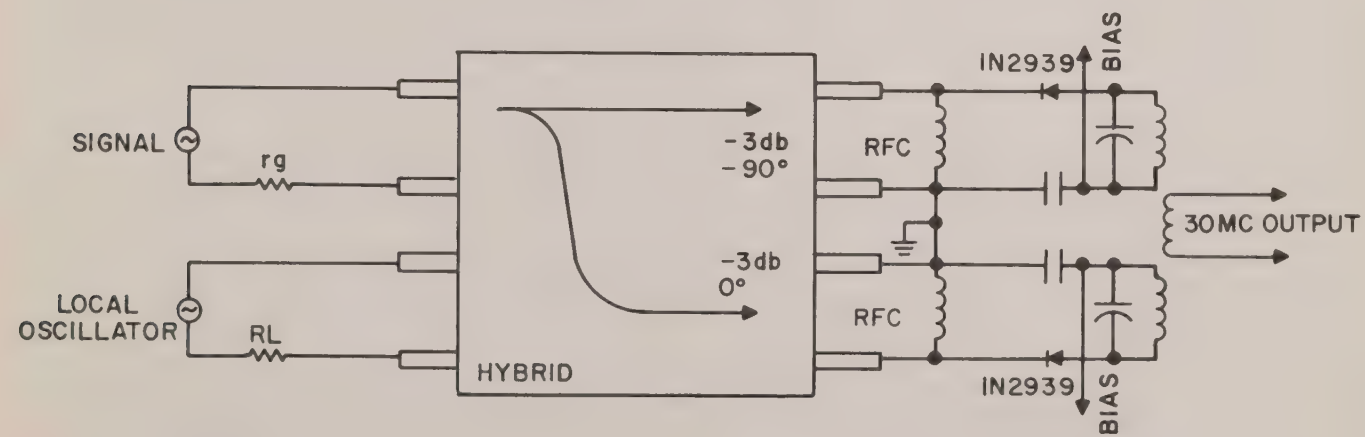

FIGURE #1

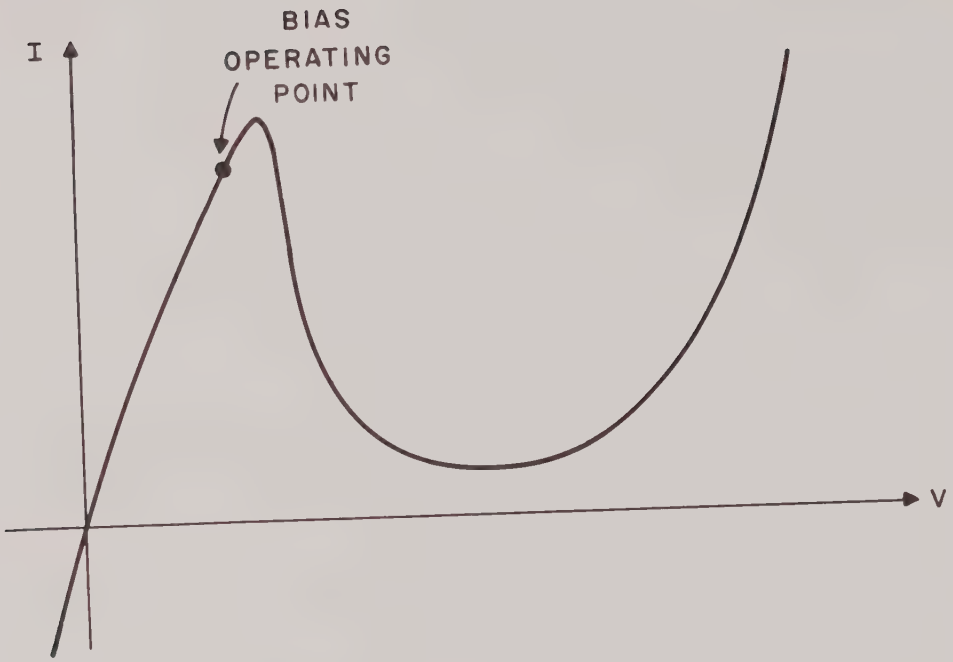

FIGURE #2
Tunnel Diode I-V Characteristic

When driven by the local oscillator the tunnel diodes and I. F. circuit appear as balanced negative conductances to the incoming signal. Hybrid action isolates the input terminal from the signal source when the local oscillator terminal is terminated in a matched load.

Measurements indicate that power gain and noise figure are critically dependent upon the power input from the local oscillator. A power input of 10 μ watts was required to yield maximum power gain and minimum noise figure. Hybrid action attenuates the local oscillator signal at the input terminal at least 20 db when the tunnel diodes are carefully matched.

Stable small signal conversion gains up to 10 db were measured at 400 mc/s. A two channel noise figure of 7 db was measured with a Hewlett Packard 342A noise figure meter at 400 mc/s. At higher frequencies power gain and noise figure deteriorate due to the cutoff frequency of the tunnel diodes that were used. With

matched tunnel diodes the down converter is both open and short circuit stable at the input terminal, without the use of a nonreciprocal device.

The author concludes that the use of Gallium Arsenide diodes would further improve noise figure, and that using a two octave hybrid should make it possible to construct a down converter with a 6:1 frequency range. He states that even in its present stage of development the hybrid coupled tunnel diode down converter offers considerable advantage over the conventional crystal mixer in the U. H. F. band.

Comments on Mr. Robertson's article :

Mr. Robertson has solved the instability problem by using a Hybrid coupler and a pair of matched tunnel diodes. He biases his diodes as Mr. Chang did in his work. (In the positive conductance region near the peak point).

His results with 2 G. E. 1N2939 are (at 400 mc) :

Conversion gain: 10 db

Two Channel Noise Figure: 7 db

(18) "MICROWAVE TUNNEL DIODE AUTODYNE RECEIVER"

by

F. Sterzer, A. Presser and A. H. Solomon
1961 International Solid State Conference-Digest of Technical Papers
pp. 88-89

Mr. Sterzer et al. describes a self-oscillating tunnel diode converter operating at a signal frequency of 1665 mc/s and a 30 mc/s I. F. Mr. Sterzer states, however, that several realizations of this circuit have been built and tested covering a range of from 200 - 2500 mc/s. The features of the converter are:

1. It lends itself readily to miniaturization, the only active circuit element required is a tunnel diode and the passive circuitry required is simple.

2. Power supply requirements are modest; a D. C. voltage of about 0.1 - 0.3 volt providings few milliwatts of power is sufficient.

3. The Autodyne can operate with unstable tunnel diodes, i. e. diodes where

$$\frac{L_D}{R_D^2 C_D} > 3,$$

(L_D = series inductance of diode - R_D = negative resistance of diode, C_D = diode junction capacitance), whereas conventional tunnel diode amplifiers (1) and non-oscillating down converters cannot. This fact is important because diodes operable at high microwave frequencies are, in general, unstable.

Figure #1 shows the circuit of the autodyne receiver. The R. F. input signal is fed through a high-pass filter (which prevents passage of the I. F. output signal to the input circuit) and through an impedance transformer into an oscillating circuit. This oscillator consists of a tunnel diode coupled to a transmission line resonator. The diode is biased by battery V_B. The value of stabilizing resistor Rs is chosen to provide a DC load line intersecting the diode current-voltage characteristic at a single point lying in the negative resistance region, i. e.

$(Rs/R_B) < R_D$. L_B is a choke to decouple the battery from the R. F. and I. F. circuits.

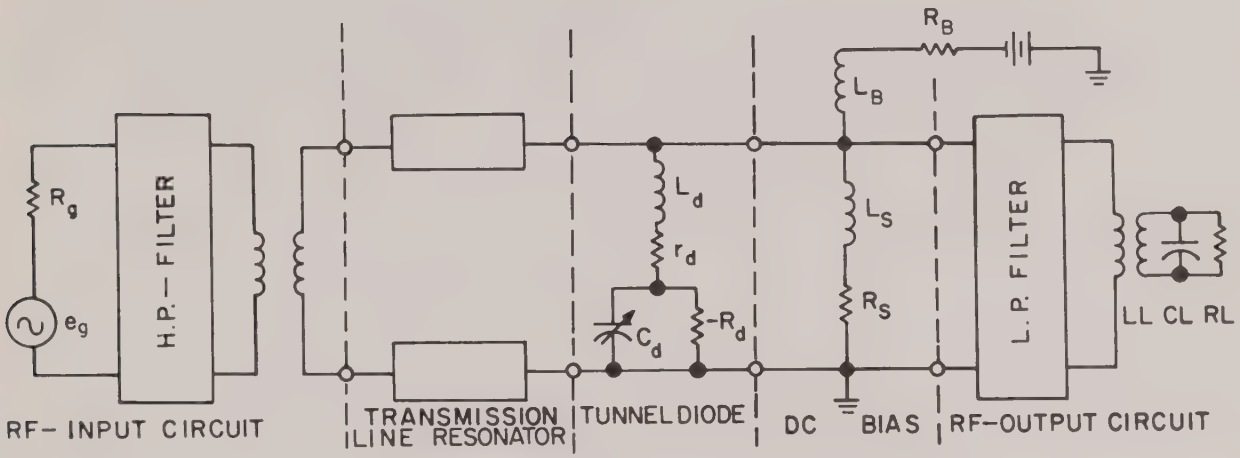

Figure #1 - Equivalent Circuit of Tunnel Diode Autodyne Receiver

The frequency of oscillation (2) is determined by the values of the tunnel diode parameters and the impedance to which the diode is coupled. To inhibit low-frequency oscillation in the D. C. circuit, the inductance of the stabilizing resistor must be chosen so that $L_S << L_B$. At the signal and oscillation frequencies, also, L_S reduces the loading due to R_S.

The nonlinearity of R_D produces mixing of signal and self-oscillation frequencies. The resultant difference frequency I. F. passes through a low pass filter and an impedance transformer to the output circuit. All other frequencies are rejected by the filter.

The output circuit consists of the load R_L, which in practice might be the input impedance of an IF amplifier, and reactive elements C_L and L_L, which tune the circuit to the desired intermediate frequency.

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(1) Smilen, L. I. and Youla, D. C., "Exact Theory and Synthesis of a Class of Tunnel Diode Amplifiers" NEC Record, 16, 1960.

(2) Nelson, D. E. and Sterzer, F. "Tunnel Diode Microwave Oscillators with milliwatt Power Output", IRE Wescon Convention Record, 1960.

Figure #2 shows conversion power gain as a function of input power. (The signal power level above which the conversion gain decreases is proportional to the peak current of the diode).

At higher signal power levels, the self-oscillation frequency is pulled by the input signal, and the intermediate frequency is reduced. This effect is illustrated in Figure #3.

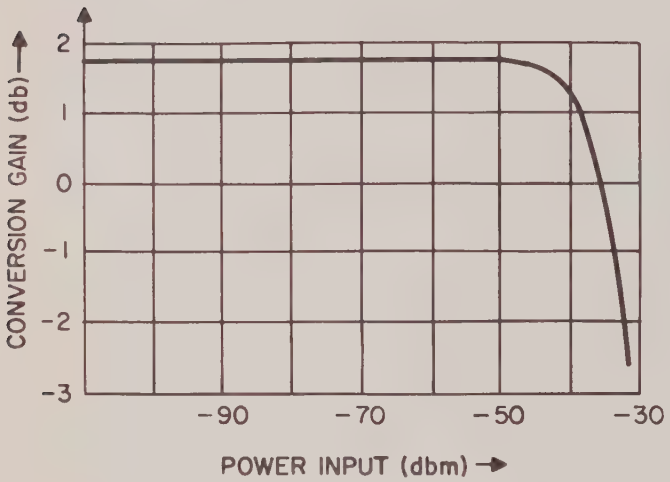

Figure #2 - Conversion Gain versus signal input power (signal frequency = 1665 Mc) for a strip transmission - line autodyne receiver. A tunnel diode with 5 ma peak current was used.

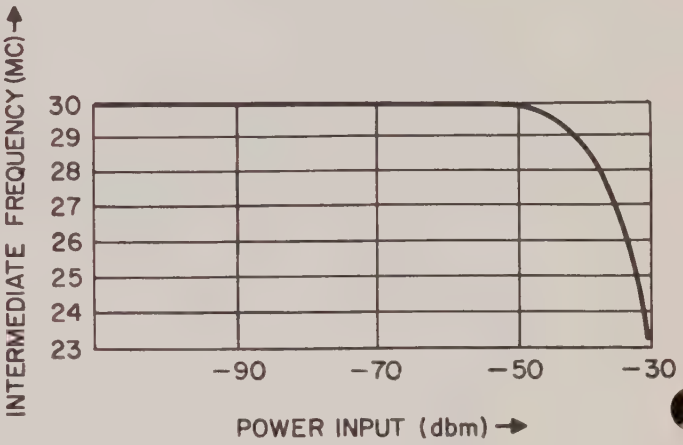

Figure #3 - Output frequency versus signal input power (signal frequency = 1665 Mc) for the autodyne receiver referred to in Figure #2.

In some autodyne circuits, responses were obtained by mixing with harmonics of the oscillation frequency with sensitivity comparable to that of the fundamental. Measurable responses at frequencies as high as the ninth harmonic have been observed.

Comments on the paper of Mr. Sterzer's et al:

Mr. Sterzer et al. gives a practical treatment of the autodyne converter in his paper and shows some very interesting results. His performance data (seemingly using a 5 ma tunnel diode) shows a conversion gain of close to +2 db for power input levels up to -50 db m and +1 db at -40 db. At the -50 db m level the oscillator starts to pull so that at -40 db m the resulting I. F. is down to 29 mc/s (from its original 30 mc/s). The signal power level above which the conversion gain decreases is said to be proportional to the peak current of the tunnel diode. It might be implied therefore that the higher the peak current of the tunnel diode the less frequency pull might be expected at high signal levels.

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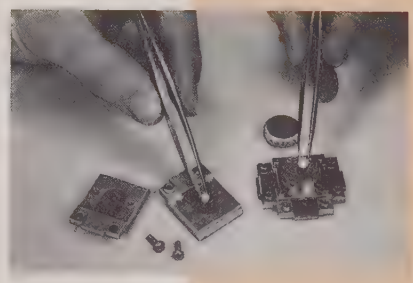

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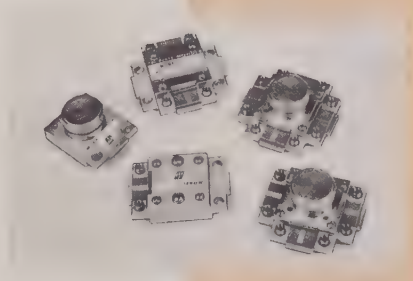

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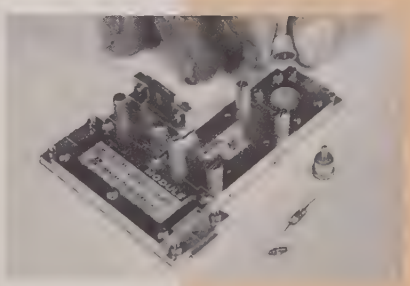

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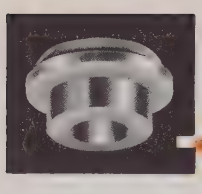
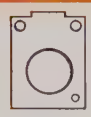
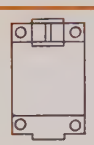
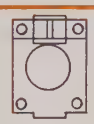
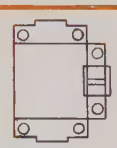
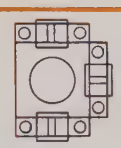
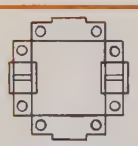
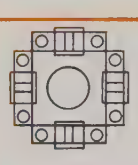


TRI-PLATE® MODULE SEMICONDUCTOR MOUNTS FOR G.E. PILL TUNNEL DIODES

TYPE	PLAIN	WITH BIAS TERMINAL	WITH GROUND RETURN	WITH BIAS AND GROUND RETURN	OUTLINE DRAWING
ONE PORT					
50 OHMS	MX77	MX77B			
10 OHMS	MX77D	MX77BD			
TWO PORT, SERIES					
50 OHMS	MX78	MX78B	MX78G	MX78K	
10 OHMS	MX78D	MX78BD	MX78GD	MX78KD	
TWO PORT, SHUNT					
50 OHMS	MX79	MX79B			
10 OHMS	MX79D	MX79BD			
THREE PORT, SERIES					
50 OHMS	MX80	MX80B	MX80G	MX80K	
10 OHMS	MX80D	MX80BD	MX80GD	MX80KD	
THREE PORT, SHUNT					
50 OHMS	MX81	MX81B			
10 OHMS	MX81D	MX81BD			
FOUR PORT, SERIES					
50 OHMS	MX82		MX82G		
10 OHMS	MX82D		MX82GD		
FOUR PORT, SHUNT					
50 OHMS	MX83				
10 OHMS	MX83D				

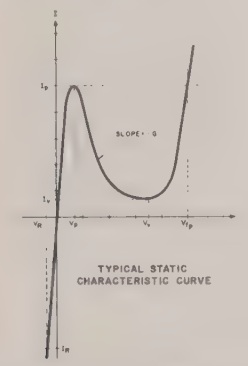
SPECIFICATIONS

1N3218, 1N3218A
Microwave Tunnel Diodes

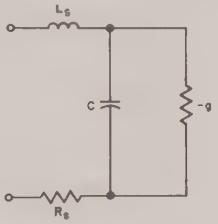
The 1N3218 and 1N3218A are germanium tunnel diodes which make use of the quantum mechanical tunneling phenomenon thereby attaining unique negative conductance characteristics and very high frequency performance. These small stripline type packages are designed for microwave communications, radar, very high frequency amplifiers and oscillator applications. The very low series inductance plus controlled low capacity permits very high frequency performance in the S band.

ABSOLUTE MAXIMUM RATINGS

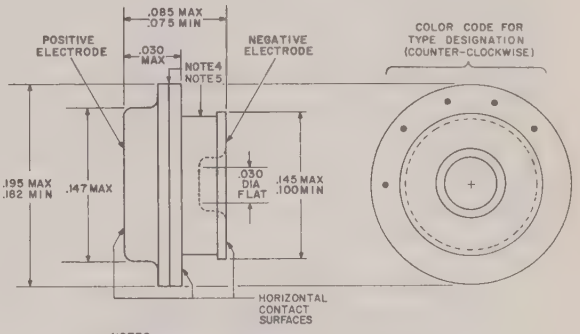
Forward Current (-55 to +100°C)	5 ma
Reverse Current (-55 to +100°C)	10 ma
Storage Temperature	-55 to +100°C



TYPICAL STATIC CHARACTERISTIC CURVE



EQUIVALENT CIRCUIT (BIASED IN NEGATIVE CONDUCTANCE REGION)



NOTES:
1. ALL DIMENSIONS IN INCHES.
2. DIMENSIONS ARE REFERENCES UNLESS TOLERANCED.
3. CONTACT SURFACES TINNED .0003.
4. WELD FLASH ALLOWED (THIS IS NOT A CONTACT SURFACE).
5. INSULATION — DO NOT APPLY CLAMPS.

ELECTRICAL CHARACTERISTICS (25°C)

		Min.	Typ.	Max.	Units
Peak Point Current	I_P	0.9	1.0	1.1	ma
Valley Point Current	I_V		0.13	0.22	ma
Peak Point Voltage	V_P		60		mv
Valley Point Voltage	V_V		350		mv
Forward Voltage ($I_F = I_P = 1$ ma)	V_{fp}	400	500	600	mv
Reverse Voltage (I_R = 1 ma)	V_R		20		mv
Negative Conductance	-G	5	9	12	$X10^{-3}$mho
Series Resistance	R_S		1.5	4.0	ohm
Series Inductance	L_S		0.3	0.5	nh
Total Capacity (1N3218)	C	1.5	7	10	pf
Total Capacity (1N3218A)	C	1.5	4	5	pf

Self-Resonant Frequency (f_{xo}) (1N3218A) $= \frac{1}{2\pi}\sqrt{\frac{1}{L_sC} - \left(\frac{g}{C}\right)^2} = 4.6$ Kmc

Resistive Cutoff Frequency (f_{ro}) (1N3218A) $= \frac{1}{2\pi}\,\frac{g}{C}\sqrt{\frac{1}{R_sg} - 1} = 3$ Kmc

Color Code: (counter clockwise on flange) 1N3218 Orange - Red - Brown - Gray - Black
1N3218A Orange - Red - Brown - Gray - Brown

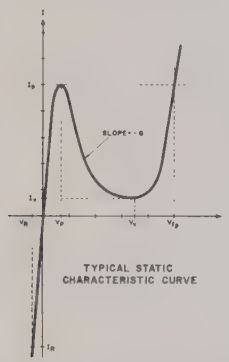
SPECIFICATIONS

1N3219, 1N3219A
Microwave Tunnel Diodes

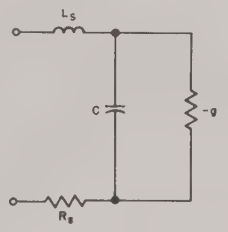
The 1N3219 and 1N3219A are germanium tunnel diodes which make use of the quantum mechanical tunneling phenomenon thereby attaining unique negative conductance characteristics and very high frequency performance. These small stripline type packages are designed for microwave communications, radar, very high frequency amplifiers and oscillator applications. The very low series inductance plus controlled low capacity permits very high frequency performance in the S band.

ABSOLUTE MAXIMUM RATINGS

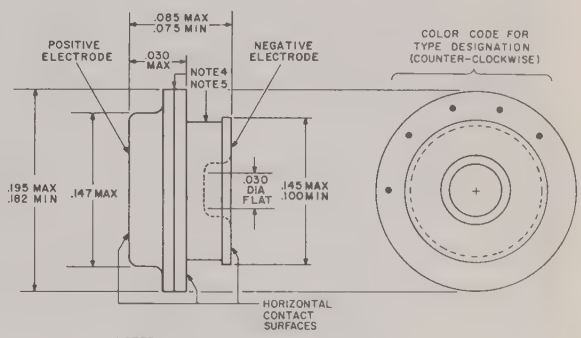
Forward Current (-55 to +100°C)	5 ma
Reverse Current (-55 to +100°C)	10 ma
Storage Temperature	-55 to +100°C



TYPICAL STATIC CHARACTERISTIC CURVE



EQUIVALENT CIRCUIT (BIASED IN NEGATIVE CONDUCTANCE REGION)



NOTES:
1. ALL DIMENSIONS IN INCHES.
2. DIMENSIONS ARE REFERENCES UNLESS TOLERANCED.
3. CONTACT SURFACES TINNED .0003.
4. WELD FLASH ALLOWED (THIS IS NOT A CONTACT SURFACE).
5. INSULATION — DO NOT APPLY CLAMPS.

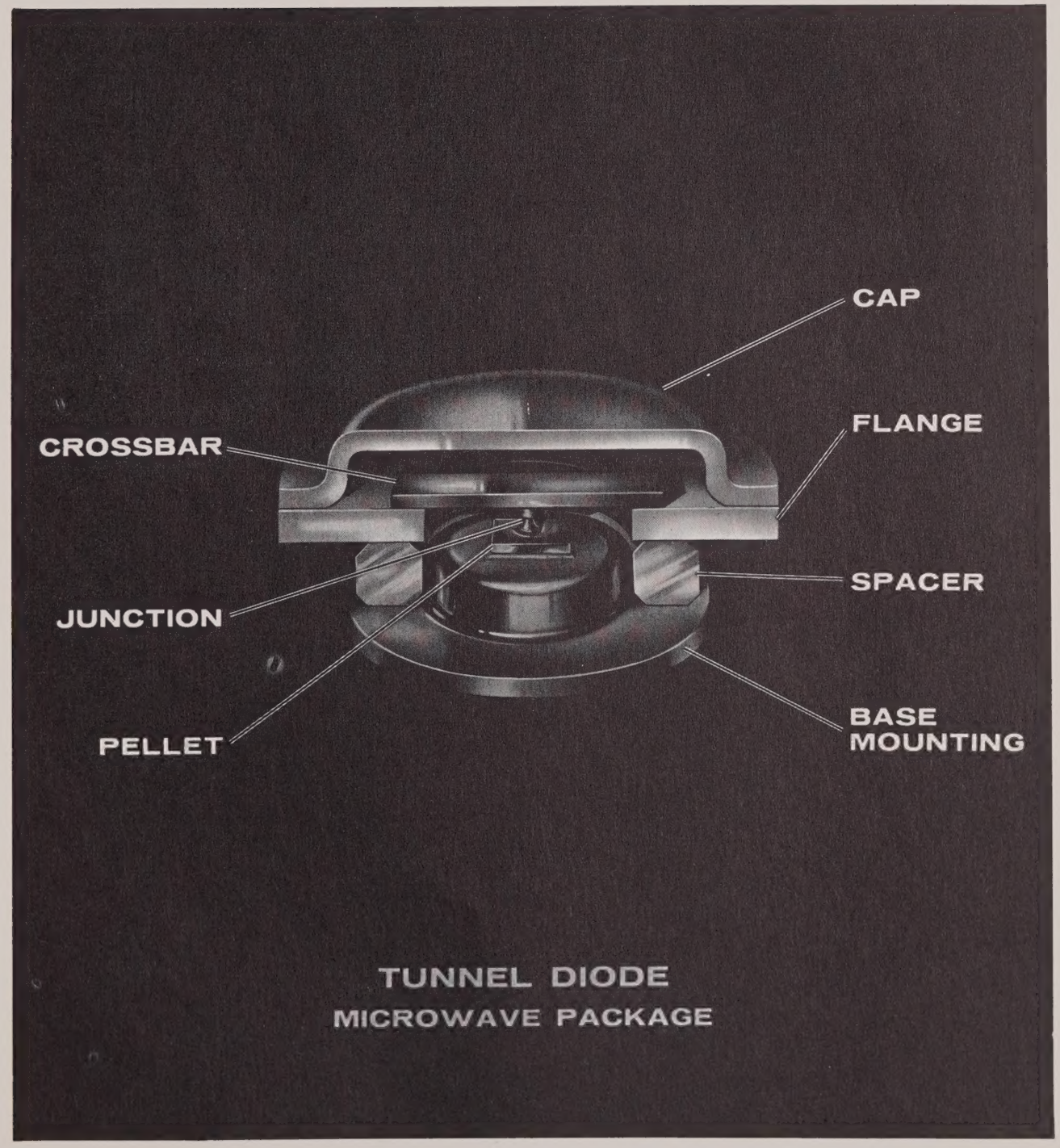
ELECTRICAL CHARACTERISTICS (25°C)

		Min.	Typ.	Max.	Units
Peak Point Current	I_P	2.0	2.2	2.4	ma
Valley Point Current	I_V		.28	0.48	ma
Peak Point Voltage	V_P		60		mv
Valley Point Voltage	V_V		350		mv
Forward Voltage ($I_F = I_P = 2.2$ ma)	V_{fp}	450	500	600	mv
Reverse Voltage ($I_R = 2.2$ ma)	V_R		20		mv
Negative Conductance	-G	10	18	25	$X10^{-3}$ mho
Series Resistance	R_S		0.7	3.0	ohm
Series Inductance	L_S		0.3	0.5	nh
Total Capacity (1N3219)	C	1.5	14	20	pf
Total Capacity (1N3219A)	C	1.5	7	10	pf

Self-Resonant Frequency (f_{xo}) (1N3219A) $= \frac{1}{2\pi}\sqrt{\frac{1}{L_sC} - \left(\frac{g}{C}\right)^2} = 3.5$ Kmc

Resistive Cutoff Frequency (f_{ro}) (1N3219A) $= \frac{1}{2\pi}\frac{g}{C}\sqrt{\frac{1}{R_sg} - 1} = 3.6$ Kmc

Color Code: (counter clockwise on flange)
1N3219 Orange - Red - Brown - White - Black
1N3219A Orange - Red - Brown - White - Brown



TUNNEL DIODE
MICROWAVE PACKAGE

MP MASTER
MA32643